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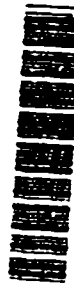
## HIGH FREQUENCY COMMUNICATIONS UNDER ADVERSE CONDITIONS

H. M. Federhen, *Project Leader*

December 1991

*Prepared for*  
The Joint Staff

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## **PREFACE**

This study was conducted by the System Evaluation Division of the Institute for Defense Analyses in response to a request from the Joint Staff (J-6).<sup>1</sup>

The paper examines various techniques which can be used to improve the performance of HF communications systems when operating in the presence of noise or jamming. Most of the discussion is devoted to directional antennas and spread-spectrum techniques.

The reviewers for the study were Dr. David L. Randall, Director System Evaluation Division (Chairman), Dr. C.L. Golliday, Mr. Harold Cheilek, and Dr. Herman Blasbalg. The author gratefully acknowledges the guidance and support of the IDA Technical Review Committee.

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<sup>1</sup> *Interoperability of U.S. and NATO Anti-Jam Communications Systems*, Contract MDA 903 84-C-0031, Task T-11-357, August 1985 and Amendments 1-4, UNCLASSIFIED.

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## PART 1

### INTRODUCTION AND SUMMARY

#### INTRODUCTION

Communication in the high-frequency (HF) band suffers from two significant problems: due to the variability of the medium (especially the ionosphere) circuits are hard to establish and maintain and, due to the relatively high noise levels, circuit quality is not as good as that provided by satellites, land lines, or microwave relays. Despite these drawbacks, there has been a resurgence of interest among the U.S. military Services in HF communications because it offers the following advantages:

- HF channels are immediately available. They can be established while on the move, or set up as soon as a unit stops in its base location.
- The HF band is the only one which will support direct over-the-horizon links without relays of any kind. HF links can range from a few kilometers to several thousand kilometers.
- HF links can be established directly by the users, without waiting for channel assignments, as is required for satellites or other long-haul systems.
- HF equipment is commercially available and is less expensive than other long-haul systems.

There are several programs now underway to improve the performance of HF systems. Automatic Link Establishment (ALE) equipment will, without operator intervention, establish and maintain contact among two or more radio stations; new voice and data modems will improve performance by automatically adapting to the noisy and variable medium; and new ECCM features will make HF communications more resistant to natural and man-made noise and to deliberate enemy jamming.

The System Evaluation Division of IDA has been tasked<sup>1</sup> to study the development of ECCM techniques for HF communications. The study will consist of three sub-tasks:

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<sup>1</sup> *Interoperability of U.S. and Allied High Frequency/Antijam (HF/AJ) Systems*. Contract No. MDA 903-84-C-0031, Task Order T-11-357, August 1985, and Amendments 1 through 4, Unclassified.

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- Development of a preliminary HF architecture, to include identification of current and planned HF radio nets, and a description of the features required for each net.
- An analysis of alternative HF ECCM techniques.
- An analysis and comparison of HF ECCM waveforms now under development or being considered, and recommendations for a minimum set of common waveforms to satisfy DoD requirements.

The HF architecture is being addressed in a separate IDA Paper. This paper discusses the alternative ECCM techniques, and the subject of ECCM waveforms will be addressed later in a separate study.

### SUMMARY

This paper first discusses both ground-wave and sky-wave HF propagation, and then derives the equation describing the performance of a radio link when interference is present. The interference can be either natural or man-made noise, or deliberate jamming. Under these circumstances, the signal-to-interference ratio at the receiver is given by:

$$\frac{S}{I} = \frac{P_T G_{TR} G_{RT} L_{IR}}{P_I G_{IR} G_{RI} L_{TR}} \left( \frac{D_{IR}}{D_{TR}} \right)^2 B \quad (1)$$

Where:  $P_T$  = Transmitter Power

$P_I$  = Interferer (jammer) power

$G_{xx}$  = Antenna Gains. For example,  $G_{TR}$  is the gain of the transmitter antenna in the direction of the receiver,  $G_{RI}$  is the gain of the receiver antenna in the direction of the interferer, and the other terms have similar meanings.

$L_{xx}$  = Excess loss over the basic free-space loss.

$D_{xx}$  = Distances.

$B$  = Bandwidth Filter Factor ( $= B_I/B_R$ ). The ratio of the interferer bandwidth to the information bandwidth of the desired signal.

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This is largely influenced by using spread-spectrum modulation which forces the interferer (jammer) to increase its bandwidth.

Each of the terms in the equation is discussed to show how it affects system performances and, in particular, the degree to which the user or designer of an HF radio link has control over the term. This is summarized in Table 1:

Table 1. Degree of Control by User

	None	Moderate	Complete
$P_T$			✓
$P_I$	✓		
$G_{IR}$			✓
$G_{RI}$			✓
$G_{IR}$			✓
$G_{RI}$	✓		
$L_{IR}$			
$L_{TR}$			
$D_{IR}$	✓		
$D_{TR}$	✓	✓	
B	✓		✓

A few observations should be made:

- $L_{IR}$  and  $D_{IR}$ , the excess loss, and the distance from the interferer to the receiver are beyond the control of the user except to the extent that siting the receiver beyond a terrain obstacle could increase  $L_{IR}$ , and the use of artillery or other weapons could cause a jammer to move, thus increasing  $D_{IR}$ .
- $D_{TR}$ , the distance from the transmitter to the receiver is generally determined only by tactical requirements. However, the use of one or more relays can break a long link into two or more shorter links, each of which will have a significantly improved S/I.
- The bandwidth filter factor, B, offers a simple and direct approach to improving HF radio link performance. Either frequency-hopping or direct-sequence pseudonoise modulation can be used to force an enemy jammer to increase its bandwidth. Each of these modulation techniques is described, and the advantages and disadvantages of each are discussed in the paper.

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It is concluded that three promising approaches to improving the performance of HF radio links should be considered:

- Increasing the transmitter power,  $P_T$ , to the extent permitted by constraints on size and weight.
- Using directional antennas which will provide beams in the direction of the transmitter or receiver, and nulls in the direction of the interferers (jammers). Such antennas are discussed in detail in the paper.
- Using spread-spectrum modulation.

## PART 2 ANALYSIS

### A. INTRODUCTION

This paper examines the techniques which may be used to maintain satisfactory communications on HF radio links, even in the face of natural or man-made interference, or deliberate jamming. Both the ground-wave and sky-wave modes of HF operation are discussed in some detail, to establish both their strengths and their limitations.

While a number of techniques can be employed to improve HF communications under adverse conditions, the discussion in this paper focuses on the selection of antennas to provide beams and nulls in selected directions and on the use of spread-spectrum modulation to gain a bandwidth advantage (and therefore a processing gain advantage) over interfering signals. These techniques appear to be the most promising for providing a substantially improved ECCM capability for HF communications.

### B. GROUND WAVE PROPAGATION

A transmitter driving an isotropic antenna in free space will generate an electric field given by:

$$E = \frac{\sqrt{30 P_T}}{D} \quad (1)$$

where:  $E$  = rms electric field strength (Volts/meter)

$P_T$  = Radiated power of transmitter (Watts)

$D$  = Distance (meters)

The power density of the radiated signal is:

$$S = \frac{E^2}{120\pi} \quad (2)$$

where:  $S$  = Power density (Watts/meter<sup>2</sup>)

$120\pi$  = Resistance of free space (= 377 $\Omega$ )

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For a receiving antenna whose effective aperture is A, the received power will be:

$$P_R = SA \quad (3)$$

If the receiving antenna is isotropic, the effective aperture is  $\lambda^2/4\pi$ , and the received power is:

$$P_R = \frac{S\lambda^2}{4\pi} \quad (4)$$

where:  $\lambda$  = Wavelength (meters)

Thus, for the case of two isotropic antennas and a lossless medium, we have, upon substituting Equations (1) and (2) into Eq. (4):

$$P_R = \frac{P_T \lambda^2}{16\pi^2 d^2} = P_T \left( \frac{\lambda}{4\pi D} \right)^2 \quad (5)$$

From which we get the Free-Space Transmission Loss equation:

$$L_{FS} = \frac{P_T}{P_R} = \left( \frac{4\pi D}{\lambda} \right)^2 = \left( \frac{4\pi Df}{c} \right)^2 \quad (6)$$

where:  $f$  = Frequency (Hertz)

$c$  = Speed of light =  $3 \times 10^8$  meters/second

If the frequency is measured in MHz and the distance in km, the free-space transmission loss (in decibels) is given by:

$$L_{FS} \text{ (dB)} = 32.44 + 20 \log f_{\text{MHz}} + 20 \log D_{\text{km}} \quad (7)$$

Equations (1) - (7) are for the special case of a lossless transmission medium which can be closely approximated under some conditions (e.g., between two aircraft). For the case of ground-wave propagation, however, there are several sources of loss which serve to attenuate the signal:

- The earth is not flat. Beyond a certain distance<sup>2</sup> from the transmitter, the receiver will see only the small fraction of the signal that is diffracted around

---

<sup>2</sup> The standard rule of thumb is that the earth may be treated as flat out to a distance  $d = 80/f^{1/3}$  where  $d$  is in km and  $f$  in MHz. For the HF band, this "flat earth region" extends to 55 km at 3 MHz, and to 26 km at 30 MHz.

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the curvature of the spherical earth. This fraction will decrease with increasing distance.

- The atmosphere absorbs part of the signal, to a degree determined by the temperature, pressure, and (particularly) humidity. This loss is not significant at HF, but becomes extremely important at VHF and above.
- The HF ground wave clings to the surface of the earth, and the earth absorbs energy from the wave. The amount of loss depends in a complicated manner on the terrain and on the physical properties of the surface.<sup>3</sup> The loss is greatest for dry ground, and lowest for sea water.

When these losses are included, Equations (1) and (5) become

$$E = \frac{\sqrt{30 P_T}}{D \sqrt{L}} \quad (8)$$

$$P_R = P_T \left( \frac{\lambda}{4\pi D} \right)^2 \frac{1}{L} \quad (9)$$

where  $L$  is the loss factor which accounts for the excess loss over free-space propagation.

Curves showing the field strength as a function of distance and frequency for a variety of surface conditions have been published by the CCIR (Ref. 2). Three of these, for dry ground, normal ground, and sea water, are shown in Figs. 1-3. Note that:

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<sup>3</sup> For a detailed discussion of this loss, complete with graphs, tables, and computational aids, see any of the three books by Terman (References 10-12).

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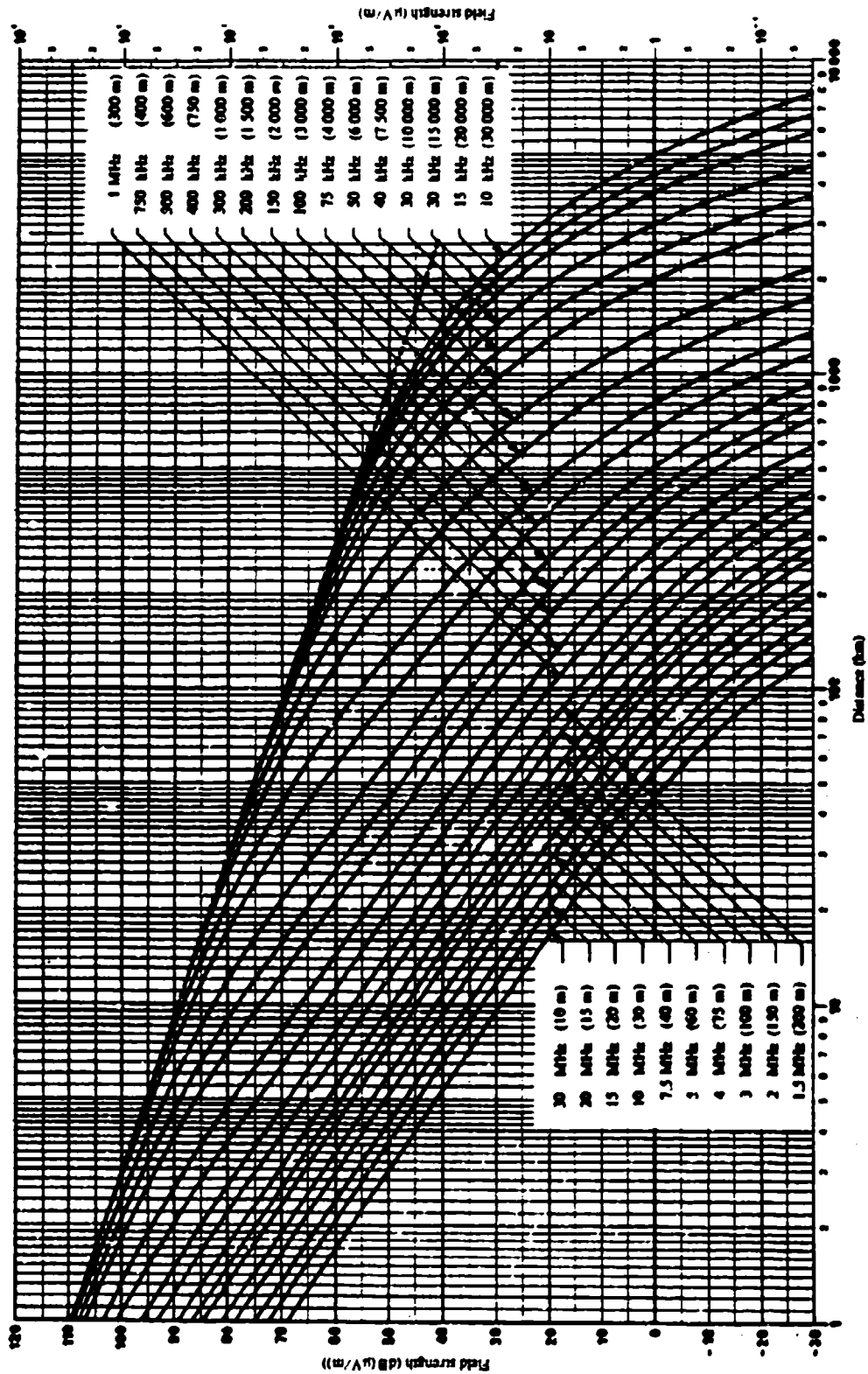


Figure 1. Ground-Wave Propagation Curves; Dry Ground,  $\sigma = 3 \times 10^{-4} \text{ S/m}$ ,  $\epsilon = 3$

(--- inverse distance curve)

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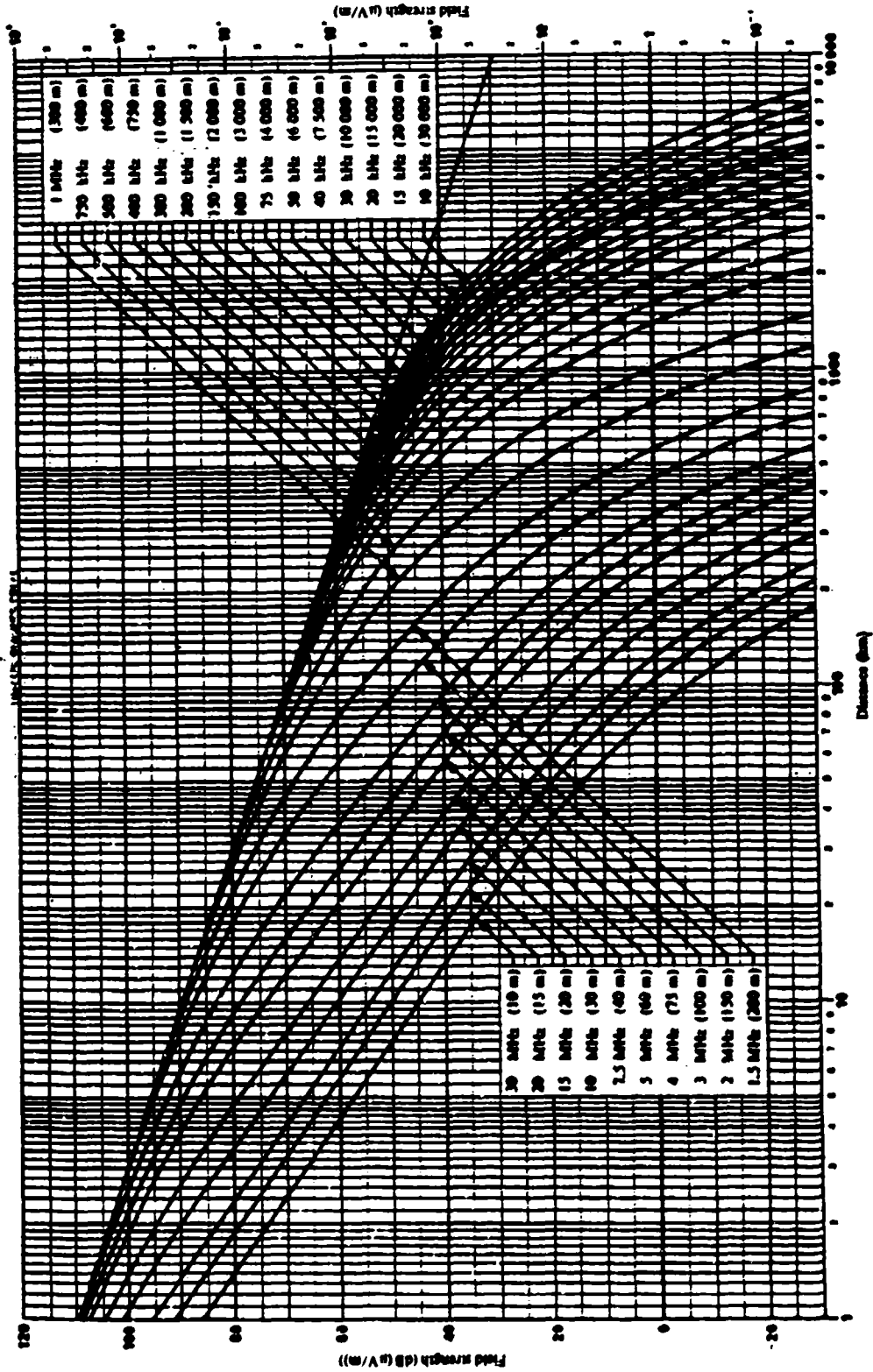


Figure 2. Ground-Wave Propagation Curves; Land,  $\sigma = 3 \times 10^{-2} \text{ S/m}$ ,  $\epsilon = 30$

(--- Inverse distance curve)

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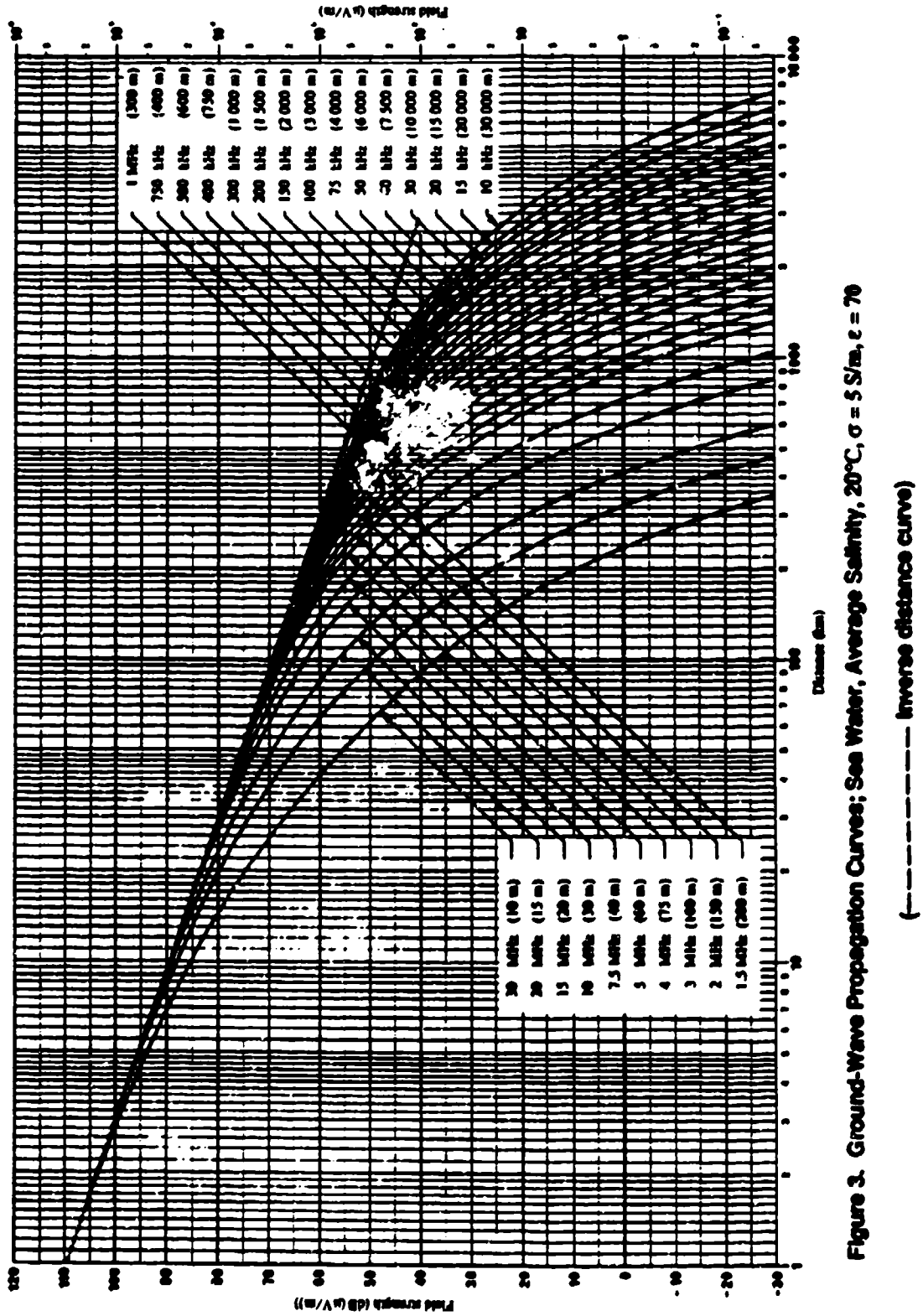


Figure 3. Ground-Wave Propagation Curves; Sea Water, Average Salinity, 20°C,  $\sigma = 5 \text{ S/m}$ ,  $\epsilon = 70$

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- The curves assume a 3 kW transmitter<sup>4</sup> feeding an omnidirectional antenna. This gives a field strength of 300 mV/m at a range of 1 km. For other values of transmitter power, the field strength at 1 km should be calculated from Eq. (1) and the scales adjusted accordingly.
- The curves assume a smooth, spherical earth. Terrain, foliage, and similar obstacles will increase the loss.
- The dashed line on each set of curves shows the free-space (lossless) field strength--i.e., the 1/d dependence given by Eq. (1).
- The right-hand ordinates of the plots show the electric field strength, E, in  $\mu\text{V/m}$ . The left-hand ordinates show  $20 \log E$  in dB relative to one  $\mu\text{V/m}$ . The latter values may be used to find the power density. Converting Eq. (2) to decibels:

$$\begin{aligned} S \text{ (dB)} &= -25.76 + 20 \log E && \text{(for E in V/m)} \\ &= -145.76 + 20 \log E && \text{(for E in } \mu\text{V/m)} \end{aligned} \quad (10)$$

It is seen that the losses increase rapidly with both frequency and distance, and that they depend strongly on the nature of the surface over which the ground wave is propagating. This is shown more clearly in Figs. 4 and 5, where the field strength is plotted as a function of frequency for fixed distances of 10 and 100 km, respectively.

The free-space loss (the 1/d loss) follows directly from the laws of physics which govern wave propagation. It is independent of the medium through which (or over which) the wave is traveling, and it cannot be changed. The values of the electric field due only to free-space loss are shown as horizontal lines at the top of Figs. 4 and 5 (89.5 dB ( $\mu\text{V/m}$ ) at 10 km and 69.5 dB ( $\mu\text{V/m}$ ) at 100 km). The additional losses due to diffraction and absorption constitute the "excess over free-space loss" and are shown schematically in the figures.

Since these excess losses depend on frequency and on the type of surface over which the wave is moving they are, to some degree, under the control of the HF system designer. A few observations can be made from Figs. 4 and 5:

---

<sup>4</sup> The curves are for a 1 kW transmitter feeding, "a short vertical monopole on the surface of a perfectly conducting earth." Since such an antenna has a gain of 3 in the horizontal plane, the effective isotropic radiated power (EIRP) is 3 kW.

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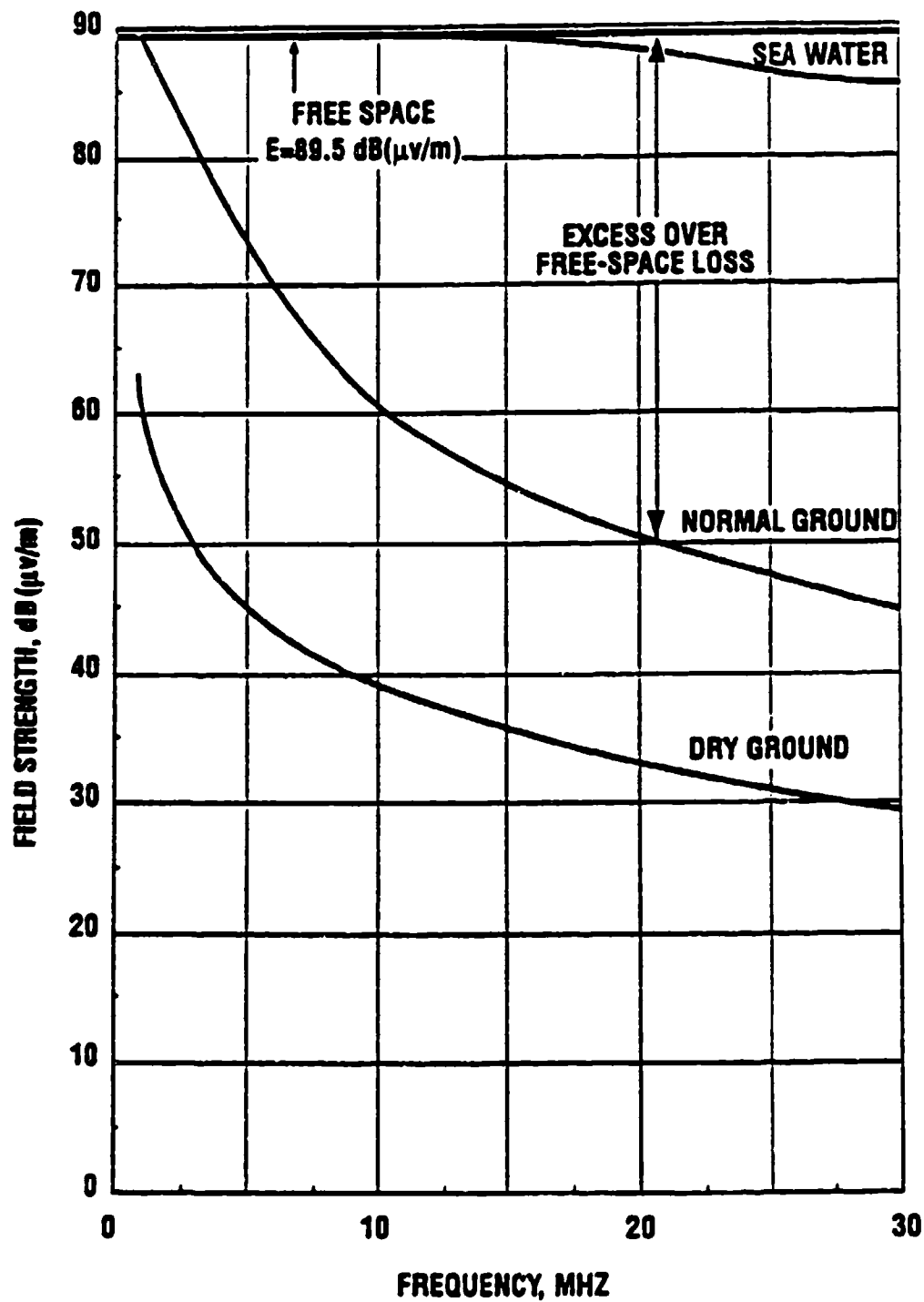


Figure 4. Field Strength at D=10 km

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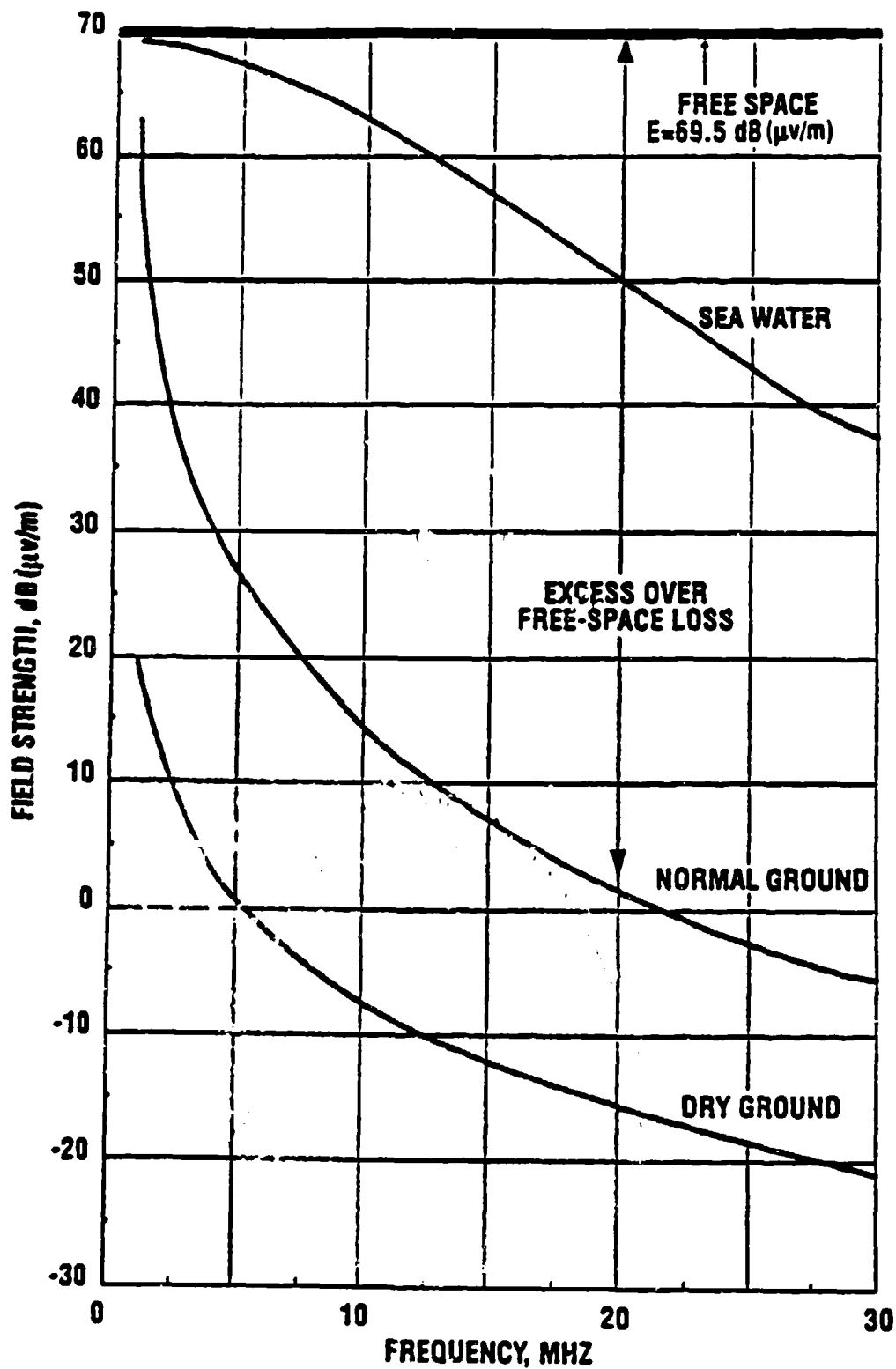


Figure 5. Field Strength at D=100 km

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- Sea water behaves like an almost perfect conductor, and has the lowest loss of all real-world surfaces. At 10 km, there is no excess loss over that of free-space for frequencies up to about 15 MHz, and less than 5 dB excess loss at 30 MHz. Beyond this distance, the excess loss does increase, but it is almost entirely caused by diffraction around the curvature of the earth rather than by absorption.
- Surfaces other than sea water have significantly higher values of excess loss, as shown in Table 2, for a distance of 10 km.

**Table 2. Excess Over Free-Space Loss at 1 km**

Surface	Frequency (MHz)		
	3	10	30
Sea water	0	0	4 dB
Fresh water	17.5 dB	28 dB	37.5 dB
Normal ground	6.5 dB	29 dB	45 dB
Dry ground	40 dB	51 dB	60.5 dB

- Figs. 4 and 5, as well as Table 2, show that minimum loss and, therefore, maximum ground wave range are achieved by operating at the lowest possible frequency.<sup>5</sup>

One other example may serve to emphasize the severe restrictions of ground wave propagation: Reliable radio reception in a quiet rural area requires a signal field strength of 1 mv/m ( $10^3$   $\mu$ v/m). Fig. 3 shows that the 3 kW (EIRP) transmitter postulated by the CCIR, operating over sea water at a frequency of 10 MHz, will provide satisfactory service out to a range of 115 km. The same transmitter operating over normal ground will have a range of only 10 km, and over dry ground the ground wave range will be less than 3 km. In order to realize the same range over normal ground as over sea water (115 km), the transmitter power would have to be increased by 50 dB--from 3 kW to 300 MW!

---

<sup>5</sup> This is one of the reasons that the commercial AM broadcast band is in the frequency band from 0.6 to 1.6 MHz.

## C. SKY WAVE PROPAGATION

The long-haul capabilities of the HF band result from the fact that, under certain circumstances, frequencies from 3 to 25 MHz may be reflected from the ionosphere, an ionized area in the earth's upper atmosphere. The ionization is largely the result of ultraviolet radiation from the sun, varies over wide limits, and is extremely hard to predict since there are not only diurnal and annual variations, but also variations with the 11-year sunspot cycle, meteor showers, and various ionospheric disturbances that occur near the poles. The ionosphere consists of several layers with differing degrees of ionization (electron density), as shown in Fig. 6.

The diurnal variation of the ionosphere is clearly seen in these diagrams. The layers and their characteristics are:<sup>6</sup>

**D Layer.** At heights from 50 to 90 km, this layer exists only during daylight hours, and the ionization density varies directly with the elevation angle of the sun. The D layer reflects VLF and LF waves, absorbs MF waves, and partially absorbs HF waves.<sup>7</sup>

**E Layer.** At a height of about 110 km, the ionization of this layer corresponds closely with the elevation angle of the sun. It reflects both MF and HF waves, and is important for HF daytime propagation at distances less than 1,500 km and for MF night time propagation at distances in excess of 150 km.

**Sporadic E Layer (not shown).** Unpredictable irregular cloud-like areas of unusually high ionization which may occur more than 50 percent of the time under certain circumstances. May absorb frequencies that would normally reach higher layers, and may also cause long-distance propagation at HF and VHF.<sup>8</sup>

**F1 Layer.** At heights of about 175 to 250 km, this layer exists only during daylight. It will occasionally reflect HF waves, but usually serves only to provide additional absorption (attenuation) to waves that are subsequently reflected by the F2 layer.

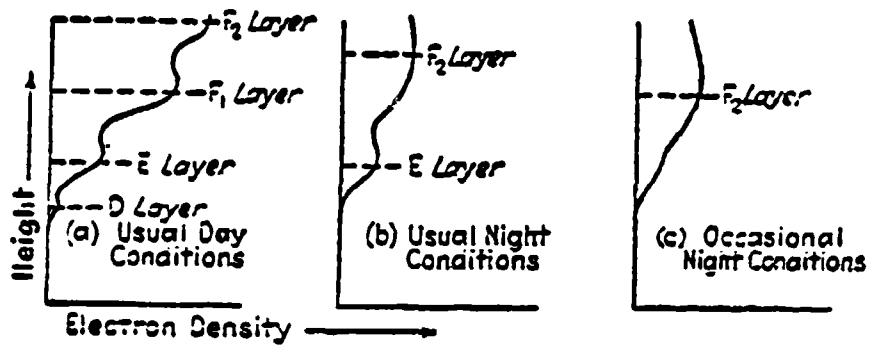
**F2 Layer.** At heights from about 250 to 400 km, this layer is the principal reflecting region for long-distance HF communication. The height and ionization vary diurnally, annually, and with the sunspot cycle, but are more constant than for the other layers. At night, the F1 and F2 layers merge at a height of about 300 km.

<sup>6</sup> Adapted from Reference Data for Engineers, Reference 6.

<sup>7</sup> The frequency bands are defined as follows: VLF = 3-30 kHz; LF = 30-300 kHz; MF = 0.3-3 MHz; HF = 3-30 MHz; VHF = 30-300 MHz.

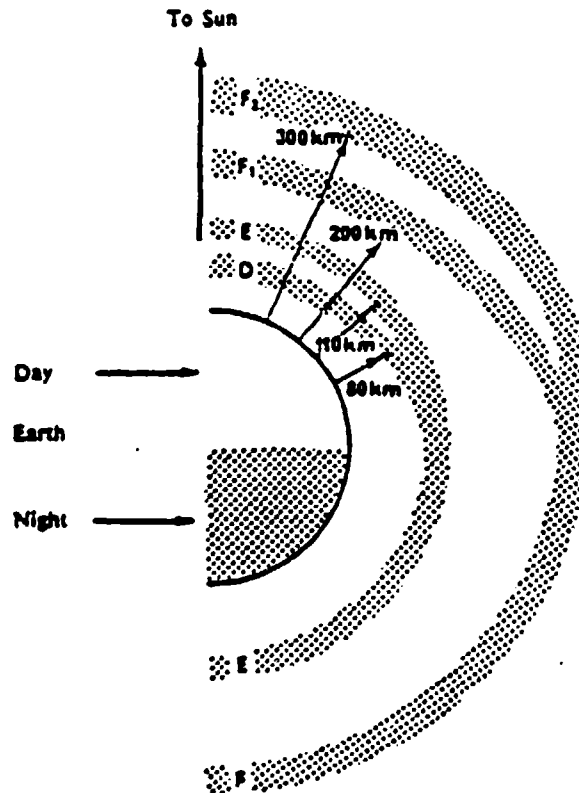
<sup>8</sup> In 1969 the author, operating an AN/VRC-46 VHF radio from his jeep in Korea, found himself in contact with a Med Evac radio net in Vietnam, at a range of more than 3,500 km. This was clearly the result of one or more reflections from very strong sporadic E layers.

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Source: Reference 10.

Figure 6a. Profiles Showing Variation of Ionization with Height



Source: Reference 3.

Figure 6b. Schematic Representation of Ionized Layers

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A radio wave incident on the ionosphere will be bent (refracted) and, if the ionization density is high enough and the frequency low enough, will be turned around and will return to earth. This is shown schematically in Fig. 7.

The radio wave actually follows a curved path in the ionosphere but, for most purposes, it may be considered as a single reflection from a surface at the virtual height  $h$ , as shown above. The degree of refraction (bending) is determined by the refractive index  $\mu$  which is given by:

$$\mu = \sqrt{1 - \frac{81N}{f^2}} \quad (11)$$

where:  $\mu$  = Refractive index

$N$  = Electron density (electrons/cm<sup>3</sup>)

$f$  = Frequency (kHz)

As shown by Eq. (11), the refractive index increases with increasing ionization ( $N$ ), and decreases with increasing frequency.

If a radio wave is launched vertically, as shown in Fig. 7 (b), there is a highest frequency at which the wave will still be reflected back to earth. This is called the critical frequency,  $f_c$ . Frequencies higher than  $f_c$  pass through the layer. For oblique incidence, the wave spends a longer time in the ionosphere, and higher frequencies will be reflected. The highest frequency which will be reflected is called the Maximum Usable Frequency (MUF) and is given by:

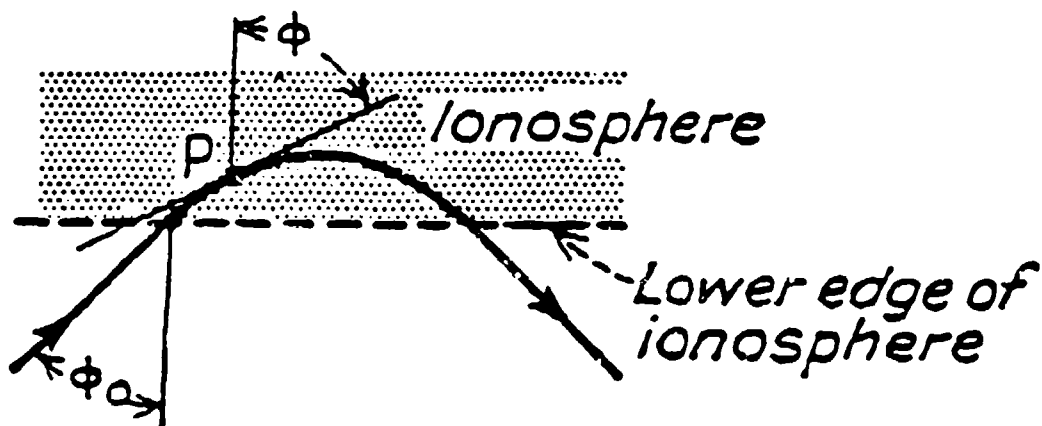
$$MUF = f_c \sec \phi_o \quad (12)$$

where:  $f_c$  = critical frequency

$\phi_o$  = angle of incidence (see Fig. 7)

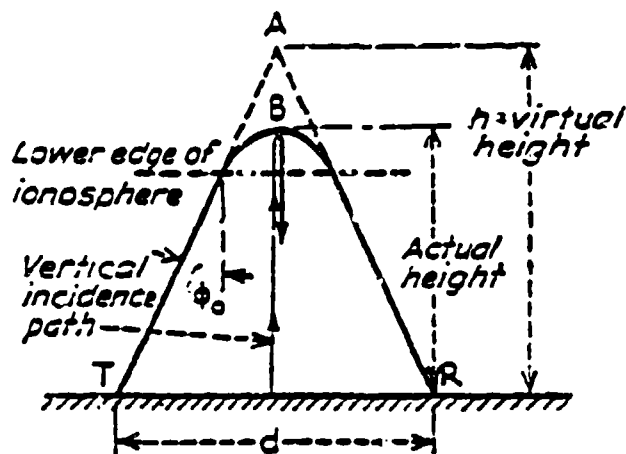
This establishes the maximum frequency at which communication can be established for a given range. It varies widely since both  $f_c$  and the height of the layer (and hence  $\phi_o$ ) vary with the time of day, season, latitude, and sunspot cycle in addition to frequent abnormal variations.

As the frequency is decreased, the ambient radio noise level and the ionospheric absorption both increase until a point is reached at which the transmitter power required for



Source: Reference 10.

Figure 7a. Refraction of Radio Wave in the Ionosphere



Source: Reference 10.

Figure 7b. Comparison of Actual Curved Path, with Single-Point Reflection at Virtual Height  $h$

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satisfactory communication becomes uneconomically high. This point is not as precisely determined as the MUF, but establishes a practical lower limit called the Lowest Useful Frequency (LUF).

Curves showing predicted values of the MUF and LUF are published regularly by the CCIR and the National Bureau of Standards (now the National Institute of Standards and Technology) among others. A typical set is shown in Fig. 8.

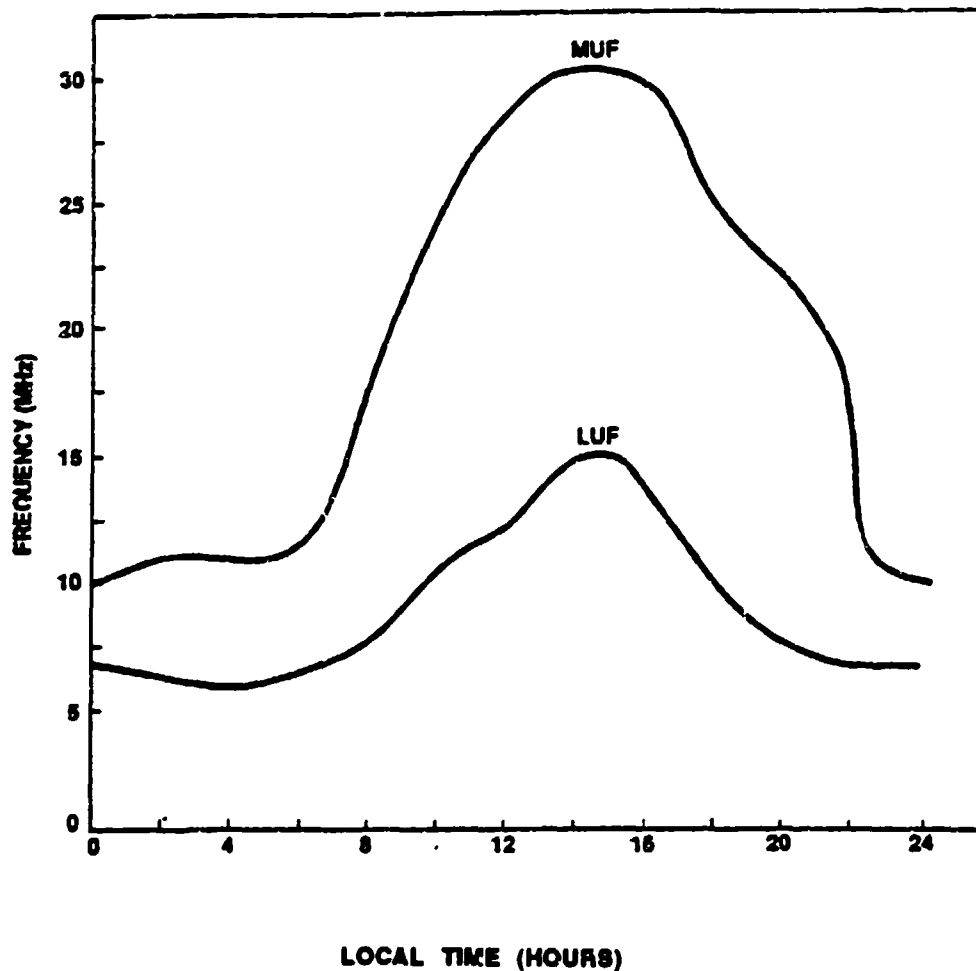


Figure 8. Typical MUF and LUF Curves

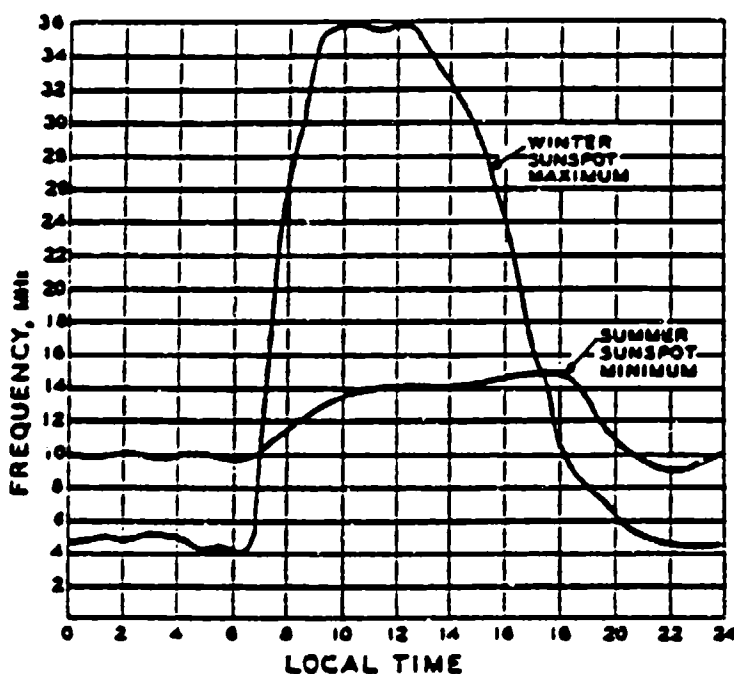
Several conclusions can be drawn from these curves:

- The range of usable sky-wave frequencies is not great during the day, and is much narrower at night.

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- At least two frequency changes per day will be required: from daytime to nighttime frequency and back again. If the band is crowded, as it usually is, additional changes may be required.
- If the daytime frequency is chosen as high as possible in order to take advantage of the lower noise levels at higher frequencies, even more frequency changes will be required.

It should also be noted that the MUF is strongly dependent on the 11-year sunspot cycle. During a sunspot maximum, when the ionospheric ionization levels are at their highest, the daytime MUF values will increase due to increased refraction, and the nighttime MUF values will decrease due to increased absorption. This is shown in Fig. 9.

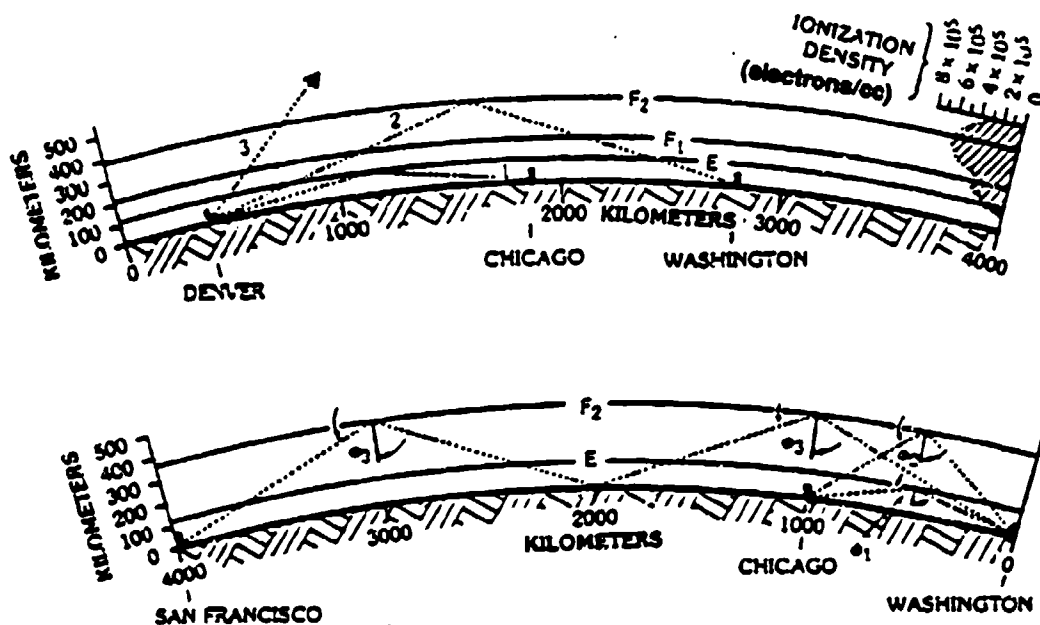


Source: Reference 9.

**Figure 9. Effect of Sunspot Activity on MUF**

At certain times, the nighttime MUF may be equal to or lower than the LUF, making HF communications difficult if not impossible.

Typical paths for sky-wave signals are shown in Fig. 10.



Source: Reference 6.

**Figure 10. Sky Wave Transmission Paths**

The upper curve shows a path from Denver to Chicago using a single reflection from the E layer, and a path from Denver to Washington using a single reflection from the F<sub>2</sub> layer. A wave launched at a higher angle or a higher frequency (path 3) will penetrate the ionosphere and be lost. Multiple reflections are possible, and the lower diagram shows a two-hop path from San Francisco to Washington using the F<sub>2</sub> layer. In all these cases, there will also be a ground wave which will extend out from the transmitter to a range of 50-100 km. The region between the end of the ground wave and the point at which the sky wave returns to earth is called the "skip distance" and no signal will be received there. If communication in this region is required, it may be possible by changing the take-off angle or the frequency of the transmitter.

The actual transmission paths are considerably more complicated than the ones shown in Fig. 10. In the first place, the height and degree of ionization of the layers will vary not only diurnally and annually, but also with latitude and longitude along the transmission path. In addition, other transmission paths are possible: three- and four-hop paths are possible although not common, and waves can become trapped between the E and

F layers and travel for considerable distances before returning to earth. These paths are essentially unpredictable and, despite all the work that has been done, HF communication remains as much an art as a science.

#### D. SIGNAL PROPAGATION AND INTERFERENCE

Eq. 9 of Section B shows the received signal power for the special case of isotropic antennas at both the transmitter and receiver. When directional antennas are used, this becomes:

$$R_S = \frac{P_T G_{TR} G_{RT}}{D_{TR}^2 L_{TR}} \left( \frac{\lambda}{4\pi} \right)^2 \quad (13)$$

where:  $R_S$  = Received signal power (Watts)

$P_T$  = Transmitter Power (Watts)

$G_{TR}$  = Gain of the transmitter antenna in the direction of the receiver

$G_{RT}$  = Gain of the receiver antenna in the direction of the transmitter

$D_{TR}$  = Distance from transmitter to receiver (meters)

$L_{TR}$  = Excess loss over ideal free-space propagation, for the path from the transmitter to the receiver.

Interference to radio reception may come from many sources: natural noise, man-made noise, or deliberate jamming.<sup>9</sup> Regardless of the nature of the interference, the result will be a reduction in the signal-to-noise (or signal-to-interference) ratio at the receiver, and the introduction of errors into the received message.

The interference power which enters the receiver and affects system performance is, by analogy to Eq. (13):

$$R_I = \frac{P_I G_{IR} G_{RI}}{D_{IR}^2 L_{IR} B} \left( \frac{\lambda}{4\pi} \right)^2 \quad (14)$$

where:  $R_I$  = Received interference power (Watts)

---

<sup>9</sup> For a discussion of noise sources, see IDA Paper P-2170, *Radio Communication in the High Frequency Band*.

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$P_I$  = Power of interferer (Watts)

$G_{IR}$  = Gain of interferer antenna in the directions of the receiver

$G_{RI}$  = Gain of receiver antenna in the direction of the interferer

$D_{IR}$  = Distance from interferer to receiver

$L_{IR}$  = Excess loss over ideal free-space propagation, for the path from the interferer to the receiver

$B$  = Bandwidth filter factor

The bandwidth filter factor  $B$  is a measure of the effectiveness of the bandpass filters in the receiver, and is a function of both the bandwidth of the bandpass filter,  $B_R$ , and the bandwidth of the interfering signal,  $B_I$ . If the entire interfering signal is included in the receiver passband, then  $B = 1$ . If the bandwidth of the interfering signal is wider than the receiver passband, then, to a first approximation,  $B = B_I/B_R$ .<sup>10</sup> This term will be particularly important when spread-spectrum ECM techniques are considered.

The signal-to-interference ratio at the receiver is then given by<sup>11</sup>

$$\frac{S}{I} = \frac{R_S}{R_I} \quad (15)$$

$$= \frac{P_T G_{TR} G_{RT} L_{IR}}{P_I G_{IR} G_{RI} L_{TR}} \left( \frac{D_{IR}}{D_{TR}} \right)^2 B \quad (16)$$

The balance of this paper will be devoted to discussing how the terms in this equation can be manipulated so as to increase the  $S/I$  ratio and thus improve the performance of the radio link.

## E. TRANSMITTER POWER

It is clear that increasing the transmitter power ( $P_T$ ) will improve the  $S/I$  ratio, and there is essentially no limit to the maximum power which can be obtained. Commercial HF

<sup>10</sup> This is valid if both the signal and interference have flat (Gaussian) spectra, and will apply to the various spread-spectrum waveforms to be considered later.

<sup>11</sup> It has been tacitly assumed that the interference is significantly stronger than natural noise sources such as thermal (KTB) or cosmic noise, since the purpose of this paper is to discuss countermeasures against interference or jamming. If the natural noise sources are included, Eq. 15 becomes  $S/I = R_S/(N + R_I)$ .

transmitters are available with output powers up to 100 kW, and OTH radar systems use large arrays of antennas and transmitters with output powers measured in megawatts. There are, however, several practical limits:

- Size and weight constrain the maximum power which can be used for certain military applications to the following limits:

Portable : < 50 W

Mobile : < 5 kW

Fixed : No limit

- The maximum efficiency of a transmitter, using the most efficient modulation (FM or single-sideband AM) is about 60%. Increasing the output power requires a corresponding increase in the size of the power source (batteries or generators), which further limits mobility.
- Increasing the transmitter power also increases the vulnerability to interception of the signal by the enemy and to possible exploitation or countermeasures. Although control of transmitter power is a standard ECCM technique, there is no point to providing power that will never be used.

## F. EXCESS LOSS TERMS

The loss terms ( $L_{IR}$  and  $L_{TR}$ ) are not generally under the control of the system engineer, since radio sites are chosen to meet tactical or strategic objectives rather than to provide low-loss transmission paths. However, the effects of diffraction and absorption should be kept in mind:

- Whenever possible, transmission paths should be chosen over water or over moist ground.
- Naval ship-to-ship or ship-to-shore HF radio links will have minimum loss and should operate reliably. Ground-to-ground links over dry sand (such as in Saudi Arabia), on the other hand, will experience high losses. This should be recognized in advance so that corrective measures, such as higher transmitter power or higher antenna gain, can be used.
- A ground-to-ground link being jammed by an airborne jammer represents the worst possible case since  $L_{TR}$  will be high and  $L_{IR}$  will be equal to 1 (i.e., no excess over free-space loss).<sup>12</sup>

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<sup>12</sup> Zero dB loss, in the form in which the equation is normally used.



## G. DISTANCE TERMS

The two distances,  $D_{IR}$  and  $D_{TR}$ , are also terms over which the system designer generally has no control, although they are very important because they appear as squared terms in Eq. (16). Thus, a 2:1 change in either of the distances yields a 4:1 change in the S/I ratio. A few general considerations might apply:

- Keep all receiver sites as far as possible from sources of man-made interference (power lines, industrial machinery, etc.).
- Take whatever steps are necessary to keep enemy jammers at the greatest possible distance.
- For marginal radio links, consider the use of a relay, thus breaking  $D_{TR}$  down into two shorter legs.

## H. ANTENNA GAIN

Of the four antenna gain terms, three can be controlled by the system designer or operator. The goal would be for  $G_{TR}$  to have a beam (lobe) in the direction of the receiver, for  $G_{RT}$  to have a beam in the direction of the transmitter, and for  $G_{RI}$  to have a null in the direction of the interferer.<sup>13</sup>

The rest of this section will discuss the characteristics of several types of HF antennas, some simple enough to be used in tactical applications and others which have better performance parameters but which would be suitable only for transportable or fixed installations. First, however, there are some considerations which apply to all antennas:

- Size. HF antennas tend to be large. The gain and size of an antenna are related by:

$$G = \frac{4\pi A_e}{\lambda^2} \quad (17)$$

where:  $G$  = Gain

$\lambda$  = Wavelength (meters)

$A_e$  = Effective area of the antenna (typically 50 to 80% of its actual physical cross-sectional area)<sup>14</sup>

<sup>13</sup>  $G_{RT}$  and  $G_{RI}$  are both provided by the receiving antenna. If the angular separation between the transmitter and the interferer is large enough ( $10^\circ$  or more), both conditions can be met.

<sup>14</sup> For thin antennas such as whips or dipoles the effective area is approximately  $\lambda L/2$ , where  $\lambda$  is the wavelength and  $L$  the length of the antenna.

Thus, to achieve a gain of 10 at a frequency of 10 MHz ( $\lambda = 30$  m),  $A_e$  must be about  $715\text{m}^2$ . The physical area must be about  $1200\text{m}^2$ , and the typical linear dimensions will be 35m.

- **Polarization.** Should be vertical for ground-wave propagation, since any horizontal component of the wave will be quickly absorbed by the surface over which it passes. For sky-wave links, either polarization may be used. The wave reflected from the ionosphere will be elliptically polarized, regardless of the polarization of the incident wave.
- **Takeoff Angle.** The angle between the main beam of the antenna and the ground should be as small as possible for ground-wave propagation. For sky-wave operation, the optimum angle depends on the length of the path and the height of the reflecting layer, as shown in Figure 11.

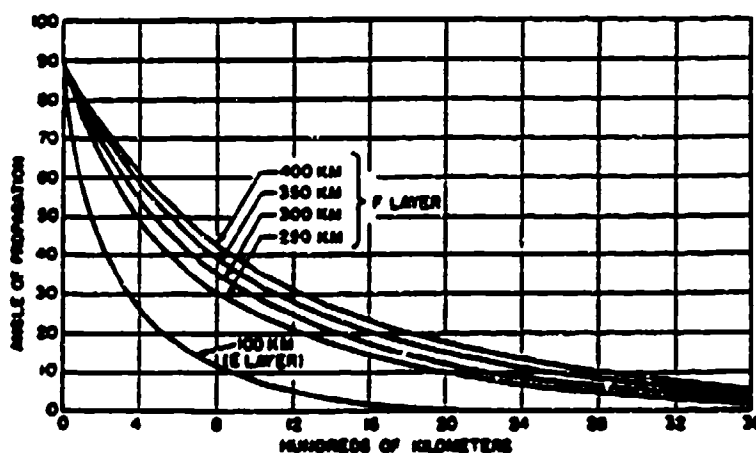


Figure 11. Vertical-Radiation Angles for One-Hop Circuits

Horizontally polarized antennas are generally used for sky-wave applications since they can be made to radiate effectively at high angles.

(1) **Whip Antennas.** The short whip is the least efficient of all the antennas to be discussed, but it is often used with portable and mobile radios since it requires no setup, and can be operated while moving. The pattern of an isolated whip is a torus centered on the whip, with a null in the axial direction. When mounted on a vehicle, the vehicle acts as a conducting ground plane and distorts the pattern. Thus, for example, a whip mounted at the left rear corner of a vehicle will have a small pattern maximum in the direction of the right front corner of the vehicle.

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For the ideal case of an antenna on a perfectly conducting ground plane, the whip has an omnidirectional pattern in the horizontal plane, and a vertical pattern as shown in Figure 12.



Source: Terman, Reference 10

Figure 12. Vertical Pattern of Whip Antenna

The corresponding gains are shown in Table 3:

Table 3. Gain of Short Antennas  
(relative to isotropic antenna)

Antenna	Gain	Gain (dB)
Isotropic in free space	1	0
Short whip	3	4.8
Quarter-wave whip	3.3	5.2

(2) Dipole Antenna. The basic dipole is a thin linear antenna fed at its center point by a balanced transmission line, as shown in Figure 13.

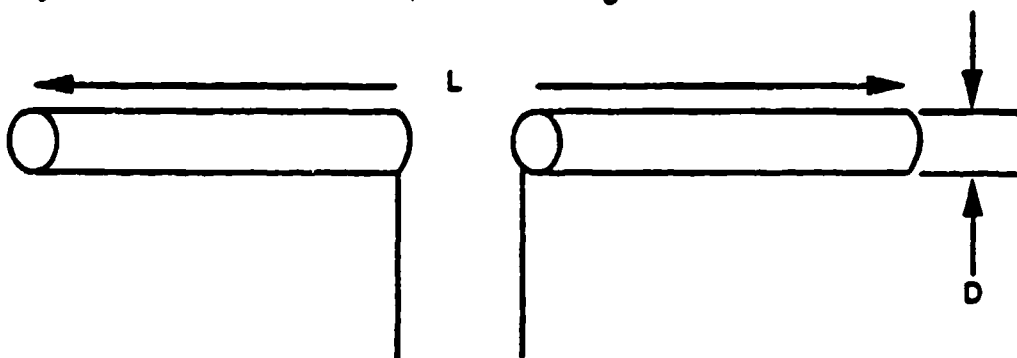
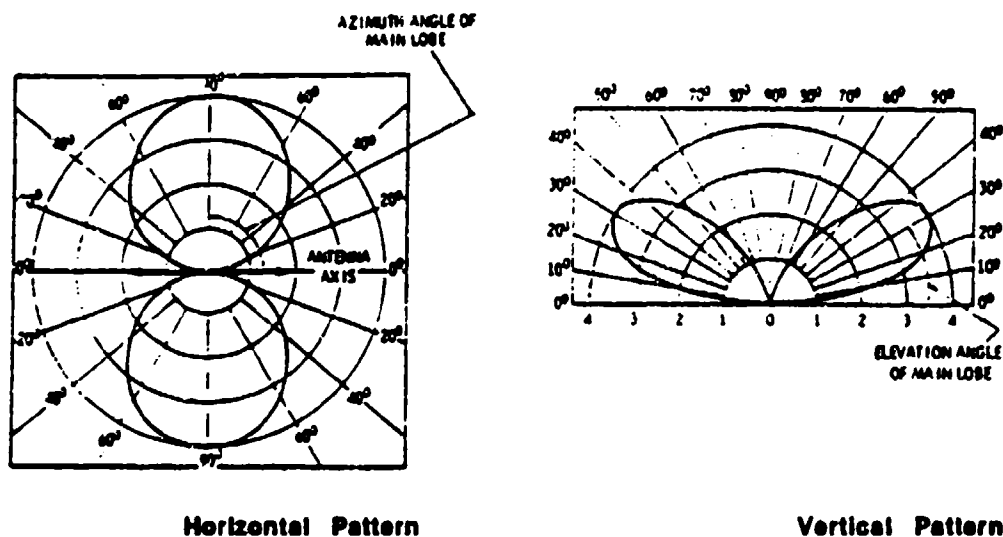


Figure 13. Basic Dipole Antenna

The antenna pattern is a strong function of the length  $L$  and of the  $L/D$  ratio.<sup>15</sup> For the present discussion we will consider only the half-wave thin-wire case ( $L = \lambda/2$ ,  $L/D = \infty$ ). For such a dipole in free space, the pattern is a torus centered on the antenna, with a null at each end in the axial direction. The gain is 1.65 (2.15 dB). The dipole may be oriented either horizontally or vertically.

If the dipole is located close to the ground, the direct radiation from the antenna and the reflected radiation from the ground interact in a fairly complicated way to create lobes and nulls in the vertical pattern.<sup>16</sup> For the case of a half-wave horizontal dipole located a half wavelength above the ground, the resultant pattern is shown in Fig. 14. The gain is now about 8 dB relative to isotropic.



Source: Reference 9.

**Figure 14. Directivity Patterns for Dipole Antenna  
One-Half Wavelength Above Perfect Ground Plane**

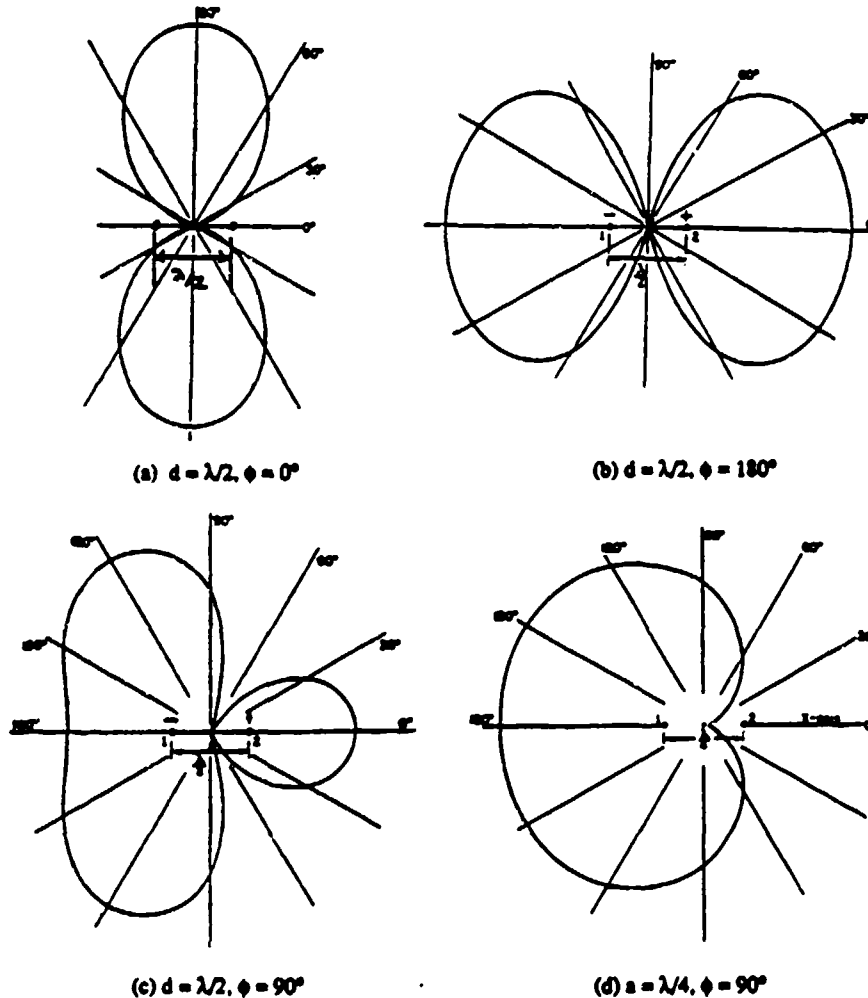
Note that there is now a null in the vertical direction, and that the pattern consists of two fairly broad beams at right angles to the axis of the antenna, each tilted upward at an angle of about  $30^\circ$ . This take-off angle is suitable for many sky-wave applications (see Fig. 11).

<sup>15</sup> See, for example, Jasik Antenna Engineering Handbook, pgs. 3-8, 3-9.

<sup>16</sup> See Jasik Antenna Engineering Handbook, pg. 21-4ff or Terman Radio Engineers Handbook, pg. 788ff.

(3) Arrays of Simple Antennas. The monopole (whip) and dipole antennas have omnidirectional radiation patterns in the plane normal to their axes, and provide no significant gain or directionality. However, arrays of these simple antennas, properly spaced and properly phased, can provide some remarkably sharp beam patterns.

For the simplest case of only two antennas, the radiation patterns<sup>17</sup> for several values of separation ( $d$ ) and phase shift between the antennas ( $\phi$ ) are shown in Fig. 15.



Source: Reference 7.

**Figure 15. Field Patterns for Pairs of Isotropic Radiators, Spaced by a Distance  $d$  and With a Phase Difference  $\phi$**

<sup>17</sup> These are actually the "array patterns" of sets of infinitesimally small isotropic radiators. The pattern of a real array is found by multiplying the array pattern by the radiation pattern of the actual antenna elements. For simple antenna elements with omnidirectional patterns (monopoles or dipoles), the actual horizon radiation pattern will be the same as the array pattern.

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The array at (a) is called a "broadside array" because its main lobes are normal to the direction of the array, (b) is called an "end-fire" array, and (d) is producing a cardioid pattern.

These simple arrays are easy to construct: two vertical guyed masts, each a quarter-wave tall, insulated from the ground, and fed at the bottom end. The patterns are such that, under many circumstances, they can be used to provide modest gain in the desired direction and/or a fairly sharp and deep null in the direction of an interferer.

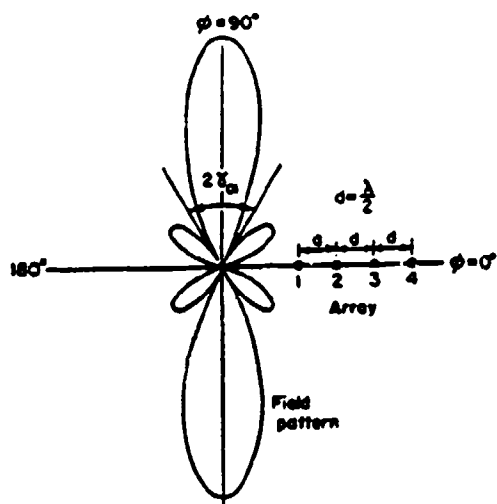
The beam sharpness is improved by increasing the number of elements in the array. Fig. 16 (a) and (b) show the patterns for 4-element broadside and end-fire arrays, and (c) shows the patterns for broadside arrays of up to 16 elements. The array gain (i.e., the gain compared to that of a single array element) is shown in Fig. 16 (d).

Arrays of simple antennas such as these can provide very useful directivities but, as noted earlier, they tend to be large. To achieve 10 dB of gain, an array must be about five wavelengths long. Such a broadside array at 10 MHz ( $\lambda = 30$  m) would consist of eleven towers, each 7.5 m tall, and extending over a length of 150 m.

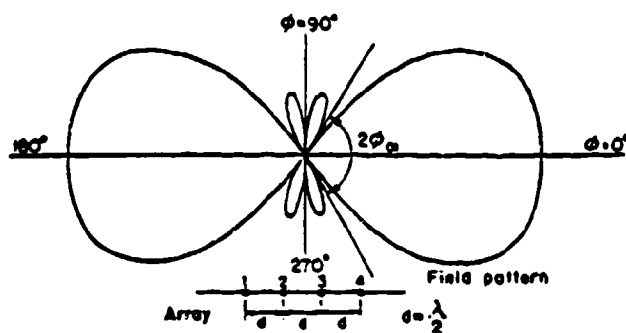
The Over-the-Horizon (OTH) radars now being built by both the United States and the USSR use arrays of simple antennas like those described here, but carried to truly impressive extremes. As an example, consider the transmit antenna of the radar site at Moscow Air Force Station in Maine. The entire antenna is 1106 m long, and varies in height from 10 to 41 m. It is divided into six sub-arrays, each tuned to a separate portion of the 5-28 MHz band. Each sub-array consists of 12 dipoles, spaced at half-wave intervals, and driven in-phase to provide a broadside radiation pattern with a beamwidth of  $7.5^\circ$ . A buried metallic ground screen extends for 230 m in front of the array to provide the equivalent of a "perfectly conducting earth," and a reflecting screen is placed behind the dipole array to convert the basic bidirectional pattern of the array into a single main beam. Since the phase of each antenna element can be separately controlled, the beam can be scanned over a  $60^\circ$  sector.<sup>18</sup> The receive antenna, located at Columbia Air Force Station in Maine, is even larger. The backscreen is 1517 m long by 20 m high, and there are 246 vertical monopole antenna elements, divided into three sub-arrays to cover the 5-28 MHz band. The receive beamwidth is  $2.5^\circ$  and, as with the transmit array, the beam can be scanned over a  $60^\circ$  sector.

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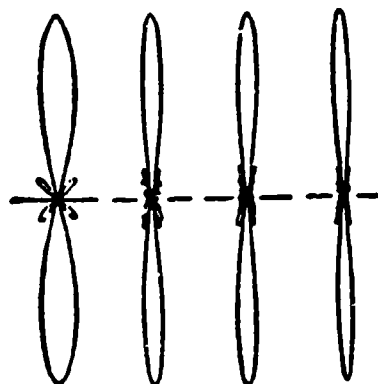
<sup>18</sup> This is a "phased array antenna," which will be discussed in detail in the next section of this paper.



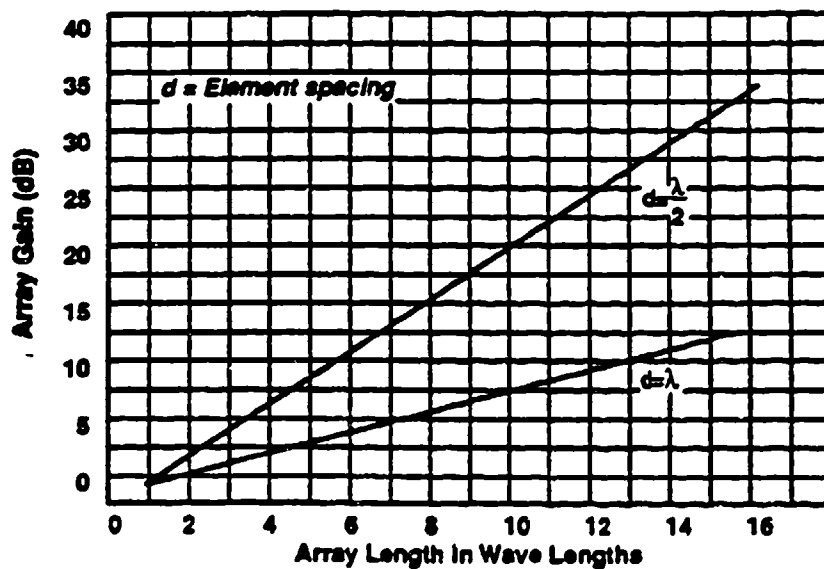
(a)  $d = \lambda/2$ ,  $\phi = 0^\circ$ ,  $n = 4$



(b)  $d = \lambda/2$ ,  $\phi = 180^\circ$ ,  $n = 4$



$n = 4 \ 8 \ 12 \ 16$   
(c)  $d = \lambda/2$ ,  $\phi = 0^\circ$



(d) Array Gain

Sources: Reference 7  
Reference 5

Figure 16. Field Patterns for Multi-Element Arrays of Isotropic Radiators

(4) Adaptive Arrays. It was shown in the last section that an array of antenna elements spaced at half-wavelength intervals and driven in-phase produces a broadside pattern, and that the same array with the elements driven  $90^\circ$  out of phase produces an end-fire pattern. It would seem logical that arrays with other element spacings and/or other phase shifts would produce beams at intermediate angles, and this is in fact the case.

Consider the general case of an array of  $N$  elements equally spaced a distance  $d$  apart. Each element has its own associated phase shifter, as shown in Fig. 17:

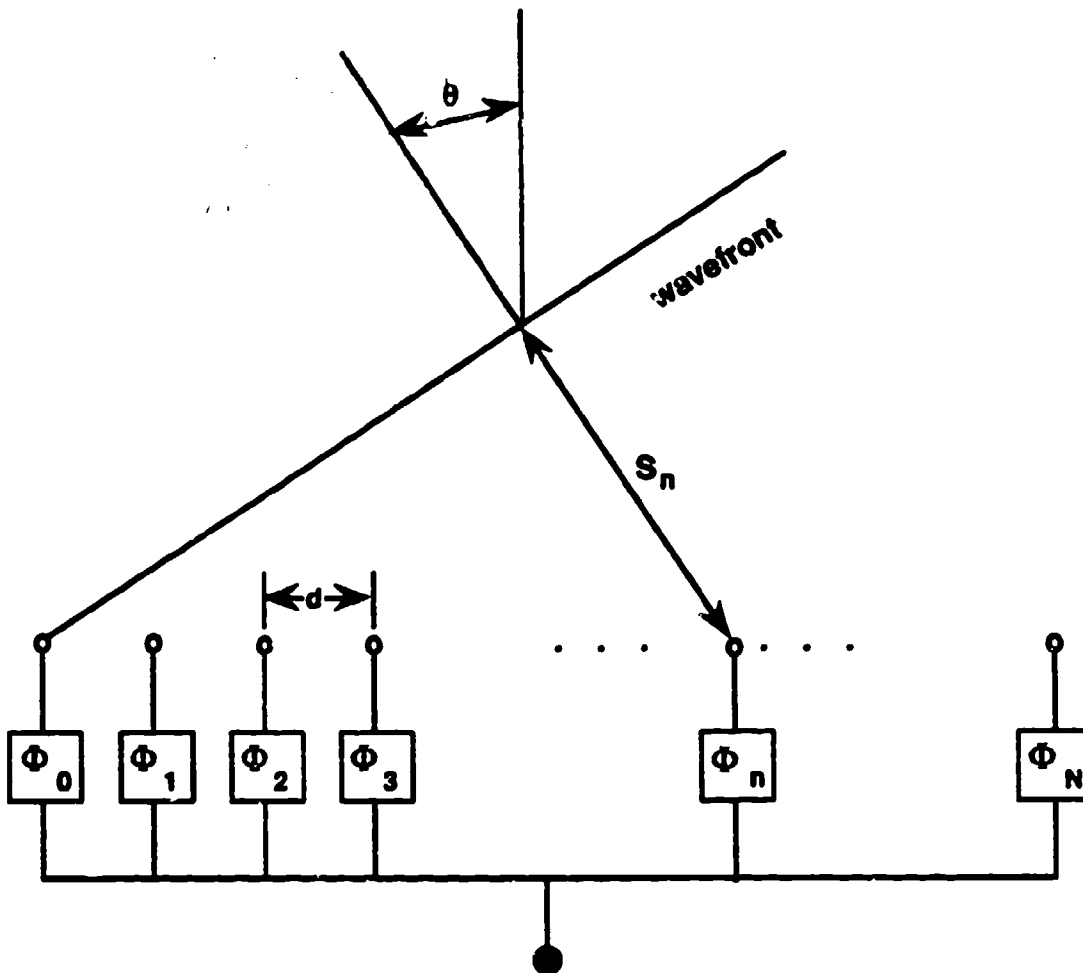


Figure 17. Linear Array of  $N$  Elements



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If this is a receiving array, and if the transmitter is very far away,<sup>19</sup> the incoming signal will appear to be a plane wave as shown in the figure. The signal arriving at the  $n^{\text{th}}$  element will be delayed with respect to the  $0^{\text{th}}$  element by a distance:

$$S_n = nd \sin \theta \quad (18)$$

or, in other words, it will be delayed in time by an amount:

$$\tau = \frac{nd}{c} \sin \theta \quad (19)$$

and will experience a phase delay given by:

$$\psi_n = \frac{2\pi nd}{\lambda} \sin \theta \quad (20)$$

To combine the signals from all the antenna elements in phase in order to produce a maximum response, the phase shifters shown in Fig. 17 should be set to:

$$\phi_n = -\psi_n = -\frac{2\pi nd}{\lambda} \sin \theta \quad (21)$$

This combination of phase shifts produces a maximum antenna response (i.e., a receive beam) at an angle  $\theta$ .

Such antennas are called *phased array* and are widely used because the beam pointing angle can be changed instantaneously by changing the settings of the phase shifters. No mechanical movement of the antenna is required.

Other modes of operation are possible:

- If every other phase shifter is shifted by  $180^\circ$  ( $\pi/2$ ), then the signals from each adjacent pair will cancel each other out, and the net output of the array will be zero. A null will have been formed in the  $\theta$  direction.

---

<sup>19</sup> The distance must be greater than  $2L^2/\lambda$ , where  $L$  ( $=Nd$ ) is the aperture of the antenna array.

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- Other combinations of phase shifter settings will form multiple simultaneous beams and/or nulls from the same array. In fact, with an N- element array, it is possible to form N-1 independent beams or nulls.

The simple approach described above will, in fact, form beams and nulls as described, but will also generate large and undesirable sidelobes. In actual arrays, both the phase and the amplitude of each element are controlled according to complex algorithms<sup>20</sup> in order to produce the desired antenna pattern with minimum sidelobes.

It should also be noted that all of the components of the array in Fig. 17 are linear and reciprocal (i.e., reversible), so that the array can be used equally well for transmitting or receiving. If, for example, the array of Fig. 17 were used as a transmitting antenna with the phase shifters set according to Eq. 21, then the signals radiated from all of the elements will add in phase to produce a transmit beam in the direction  $\theta$ .

One final capability can be added to array antennas if it is possible for the receiver to distinguish the desired signal(s) from all the other signals, interferers, and jammers that it may see.<sup>21</sup> In this case it is possible, under computer control, to set the amplitudes and phases of the antenna elements so as to maximize the received S/I ratio. The net result will be to automatically form beams in the directions of the desired signals and nulls in the directions of the major sources of interference. This is called an *adaptive array*. The performance of such an array is shown in Fig. 18.<sup>22</sup>

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<sup>20</sup> The process is called *tapering*.

<sup>21</sup> This can be accomplished by adding a pilot tone or a special code to the desired signal or, in the case of Direct-Sequence Spread Spectrum Signals (see page 54ff) by using the spreading code itself.

<sup>22</sup> Taken from Hans Steyskal "Array Error Effects in Adaptive Beam Forming," Microwave Journal, September 1991.

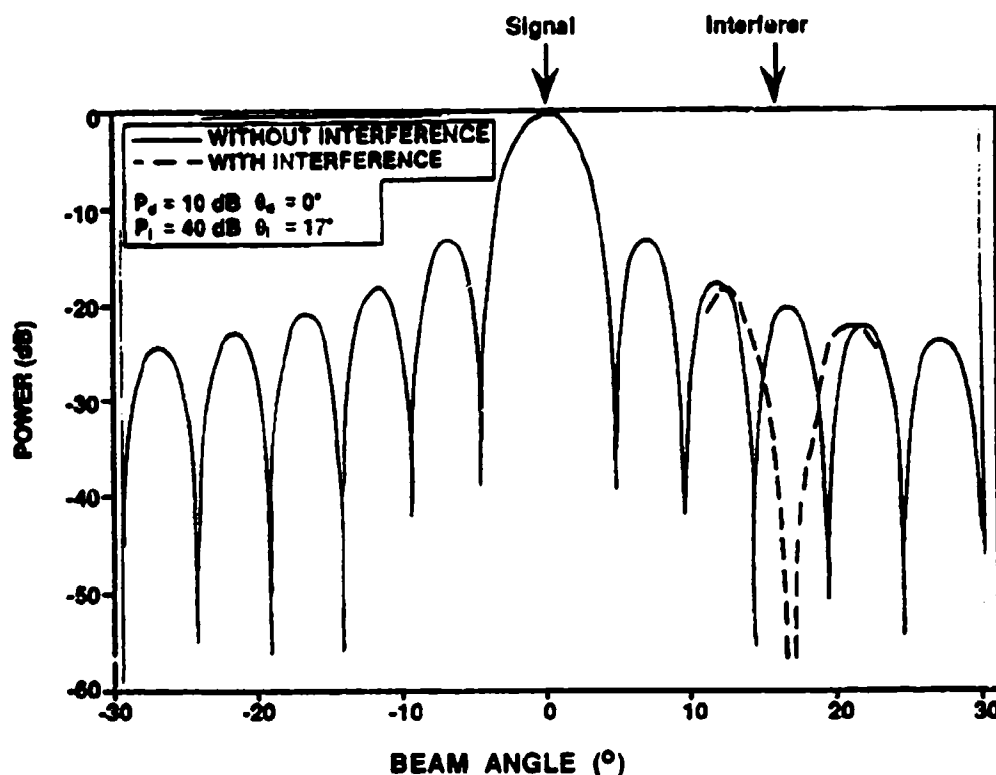


Figure 18. Normal and Adapted Antenna Patterns for a 24-Element Array with  $\lambda/2$  Spacing

In this case, the desired signal is located on the axis of the antenna at  $0^\circ$  and the interfering signal, which is 30 dB stronger, is at an angle of  $+17^\circ$ . In the normal (unadapted) mode, the interferer is at the peak of one of the sidelobes of the antenna. After adaptation, a null of more than 60 dB has been formed to cancel out the interfering signal. In practice, this process would take place instantaneously and continually, to handle multiple interferers and to adapt to motion of both the desired and interfering signals.

Both phased arrays and adaptive arrays are widely used in radar antennas since they can form multiple beams, allow instantaneous beam shifting, and can have special features such as track-while-scan. The search and track antennas on military aircraft are almost invariably phased arrays, as are the Pave Paws early-warning radars.<sup>23</sup> They may also be used for special HF systems such as OTH radars, but their use for more general applications at HF is problematical because of the size and complexity of both the phase shifters and the arrays themselves.

<sup>23</sup> And the infamous Krasnoyarsk radar.

(5) Long Wire Antennas. A resonant wire antenna is one that is an integral number of half wavelengths long, and is open (i.e., not grounded) at both ends. When such an antenna is fed from one end, standing waves will be generated along the line.<sup>24</sup> The radiation patterns, for lengths of  $\lambda/2$  to  $8\lambda$ , are shown in Fig. 19.

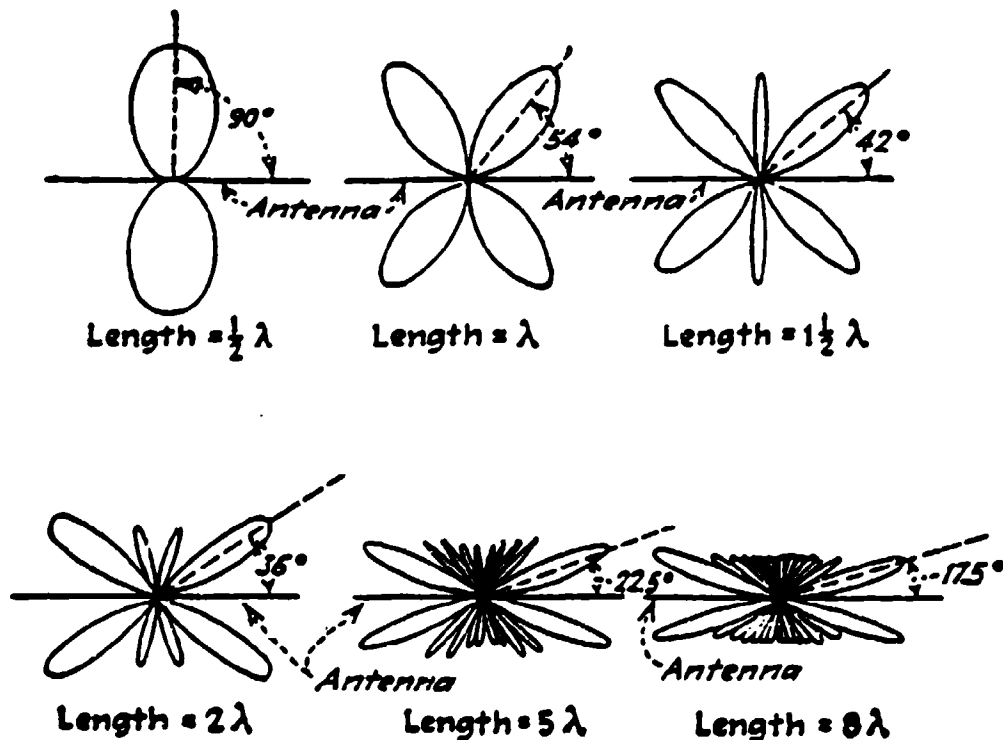
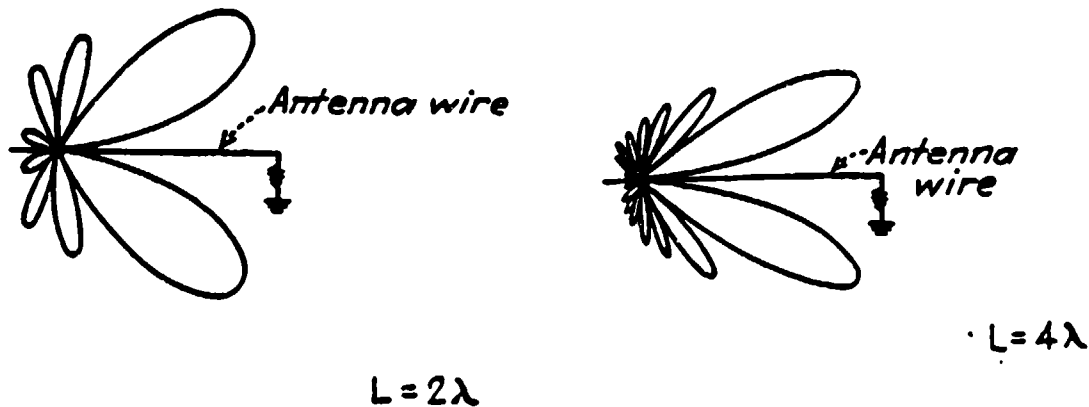


Figure 19. Field Pattern of Resonant Wire Antenna in Free Space

The actual radiation pattern is, of course, the figure of revolution of the cross section shown above about the axis of the antenna, and appears in general as a series of nested coaxial cones. The total number of conical lobes in each pattern is equal to the length of the line in half-wavelengths (i.e., 1, 2, 3, 4, 10, and 16 for the examples in Fig. 19).

If the line is terminated at one end in its characteristic resistance, waves traveling down the line will be absorbed rather than reflected, and there will be no standing waves. The line is then said to be "nonresonant" and the radiation patterns for two such lines are shown in Fig. 20.

<sup>24</sup> These result from the addition of the signal fed from one end and the wave reflected from the other end.



**Figure 20. Field Pattern of Nonresonant Wire Antenna in Free Space**

Again, the three-dimensional patterns look like nested cones, and the number of lobes in each pattern is equal to the length of the line in half-wavelengths. There are, however, several significant new features for the nonresonant long-wire antenna.

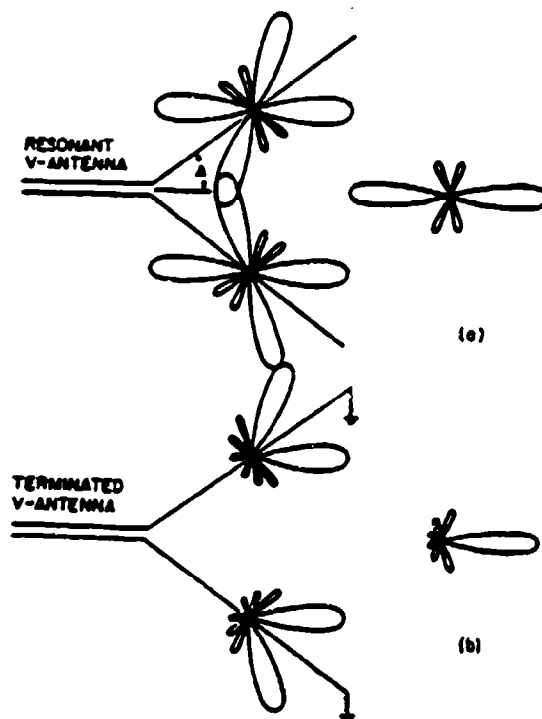
- Half of the power fed to the antenna will be dissipated in heat in the terminating resistor.
- Whereas the pattern of the resonant antenna was bidirectional, the nonresonant antenna concentrates its energy in one direction (to the right in Fig. 20).
- Since the antenna is nonresonant, the length is less critical. Although Fig. 18 shows the patterns for lengths of  $2\lambda$  and  $4\lambda$ , the patterns for intermediate lengths will not differ significantly from the ones shown. Conversely, for an antenna of fixed length, the pattern will change only slowly with frequency, and the antenna can be used over a much broader bandwidth.

Long-wire antennas are not widely used both because the conical (or biconical) patterns do not lend themselves well to either area broadcast or point-to-point applications, and because the antenna pattern is dependent on the frequency and/or wire length.

However, they also form the basis for two of the most important types of HF antennas: the V and the rhombic.

(6) **V Antennas.** If two long wire antennas are arranged to form a V with the apex angle equal to twice the angle that the first lobe in the pattern of the long wire antenna makes with the wire, then the radiation lobes on the line bisecting the V will reinforce each other, and the lobes in other directions will tend to cancel. For a two-wavelength long wire, the angle of the first lobe is  $36^\circ$  (see Fig. 19), and the resultant patterns for a V antenna with an apex angle of  $72^\circ$  is shown in Fig. 21.

Note that the resonant V antenna has a bidirectional pattern, while the terminated V has only a single main lobe. The latter pattern is preferable for most applications but, as before, half of the transmitter power is dissipated as heat in the terminating resistors.<sup>25</sup>

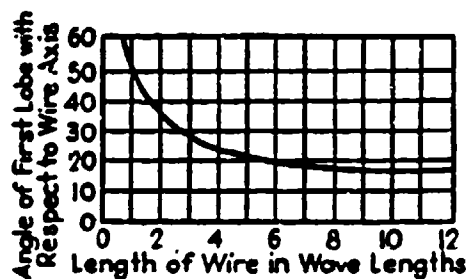


Source: Reference 5.

Figure 21. Field Patterns of (a) Resonant Two-Wavelength V Antenna; (b) Terminated Two-Wavelength V Antenna

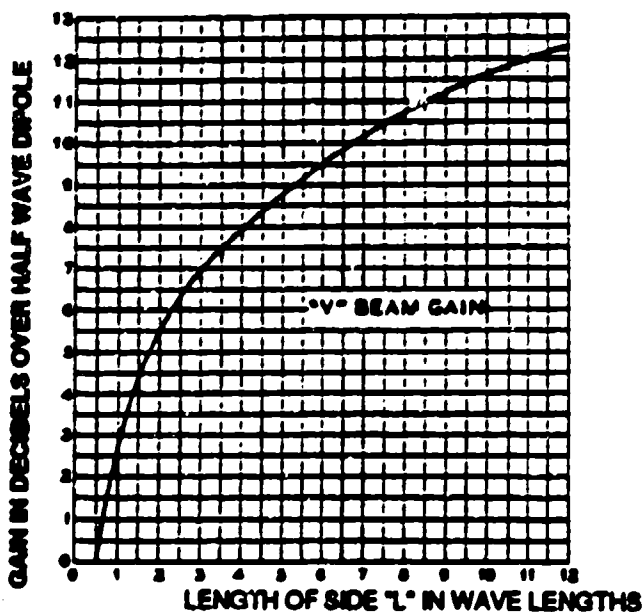
<sup>25</sup> The unidirectional pattern may also be obtained by using a second V antenna located  $\pi/4$  behind the first antenna, and feeding the two antennas  $90^\circ$  out of phase. This end-fire V array is more complicated, but saves the power which would otherwise be wasted in the terminating resistors.

As the legs of the V are made longer, the lobes of the two long-wire antennas move closer to their axes and become narrower and stronger, and the gain of the V antenna will increase. This is shown in Fig. 22 where it is seen that a five-wavelength antenna (150 m legs at a frequency of 10 MHz) has a gain of almost 9 dB over that of a half-wave dipole, or about 11 dB over an isotropic antenna.



Source: Reference 10.

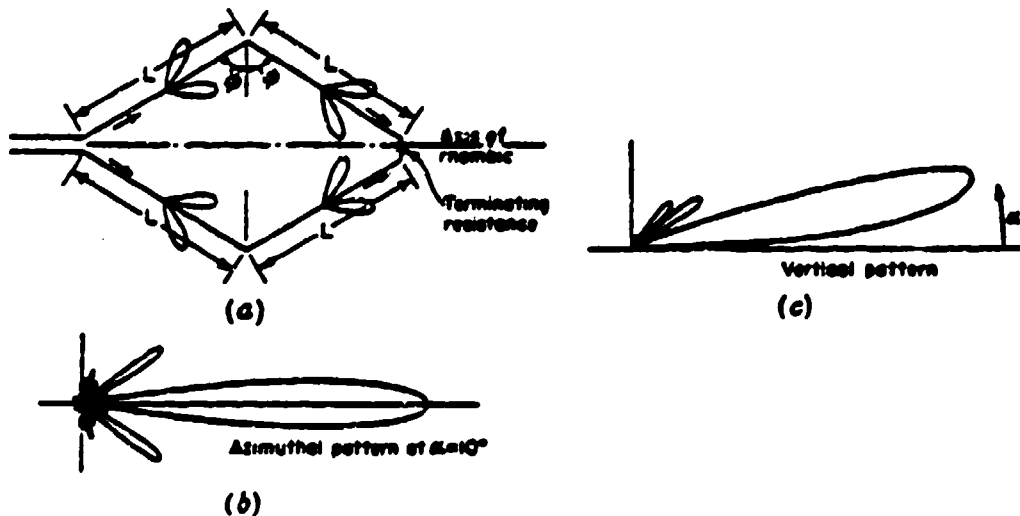
Figure 22a. Angle of First Lobe of Long-Wire Antenna



Source: Reference 10.

Figure 22b. Directive Gain of V-Beam Antenna

(7) **Rhombic Antennas.** Rhombics are widely used for long-haul point-to-point HF circuits between fixed stations because they are relatively simple to install and maintain, provide high gain at moderate cost, and can operate over a relatively wide frequency band. The rhombic antenna consists of two V antennas (or four long-wire antennas) connected as shown in Fig. 23.



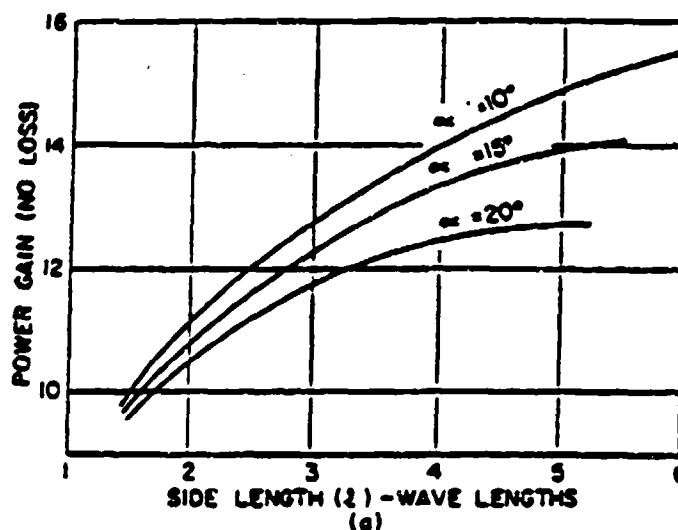
Source: Reference 7.

**Figure 23. Terminated Rhombic Antenna (a) With Azimuthal Pattern (b) and Vertical Plane Pattern (c) for a Rhombic 6 Wavelengths Long on Each Leg,  $\phi = 70^\circ$ , and at a Height of 1.1 Wavelengths Above a Perfectly Conducting Ground (After A. E. Harper)**

A single terminating resistor, located as shown, makes the antenna nonresonant. If the parameters  $L$  and  $\phi$  are chosen properly, the radiation lobes of the four long-wire antennas will add in phase to produce a single main beam along the axis of the antenna.

As with the V antenna, the main beam gets narrower and the gain of the antenna increases as the leg length,  $L$ , is increased. This is shown in Fig. 24.





Source: Reference 8.

**Figure 24. Power Gain of Free-Space Rhombic Antenna of Optimum Design for Maximum Field Intensity at Vertical Angles  $\alpha$  of 10, 15, and 20°**

The vertical take-off angle ( $\alpha$  in Fig. 24) should be chosen to be optimum for the radio link, based on both the path length and the height of the ionosphere. This is done by controlling three parameters of the rhombic antenna:<sup>26</sup>

1. As the length ( $L$ ) of the legs is increased, the lobes of the individual long-wire antennas will move closer to the axes of the wires (see Section (4) and Fig. 22).
2. If the rhombus is made narrower (i.e., if the angles  $\phi$  in Fig. 23 are increased), the conical patterns of the individual long-wire antennas will overlap. Then, instead of a single horizontal beam along the axis of the antenna, there will be two axial beams, one above and the other below the plane of the antenna. For

<sup>26</sup> For a detailed discussion of rhombic antenna design see Jasik Radio Engineering Handbook, Chapters 4 and 21; or Kraus Antennas, pg. 408ff.

proper choices of  $L$  and  $\phi$ , the beams will make the desired vertical angle ( $\alpha$ ) with the horizontal.

3. Now if the antenna is placed at the proper height ( $H$ ) above a conducting ground, the lower beam will reflect off the ground in such a way as to add in phase with the upper beam, and all the energy will be concentrated into a single beam, pointing along the axis of the rhombic and with a vertical take-off angle  $\alpha$ .

The antenna designer must choose  $L$ ,  $\phi$ , and  $H$  to give the desired value of  $\alpha$ . The formulas for doing this are given in Kraus Antennas, pgs. 410 and 411.

As the frequency is changed, the leg length  $L$  and the height  $H$ , measured in wavelengths, will also change. The angles that the lobes make with the individual wires will therefore change, but the changes in the four wires tend to cancel each other out. The net result is that the beam is broadened somewhat for frequencies other than the optimum design frequency. As a rule, a rhombic antenna will operate satisfactorily over a frequency range of 2.5 to 1.

(8) Yagi Antennas.<sup>27</sup> In all of the antennas discussed so far, all of the elements have been driven--i.e., connected directly to the transmitter. It is also possible to use parasitic (non-driven) elements which affect the antenna by virtue of the currents which are induced in them. This is the basis for the Yagi antenna. The antenna and its radiation patterns are shown in Fig. 25.

If a parasitic element is longer than the driven element, it will reflect energy back toward the driven element and is called a reflector. If it is shorter, it sends energy away from the driven element and is called a director. The three-element array shown here has a gain of about 5dB over that of a single dipole.<sup>28</sup>

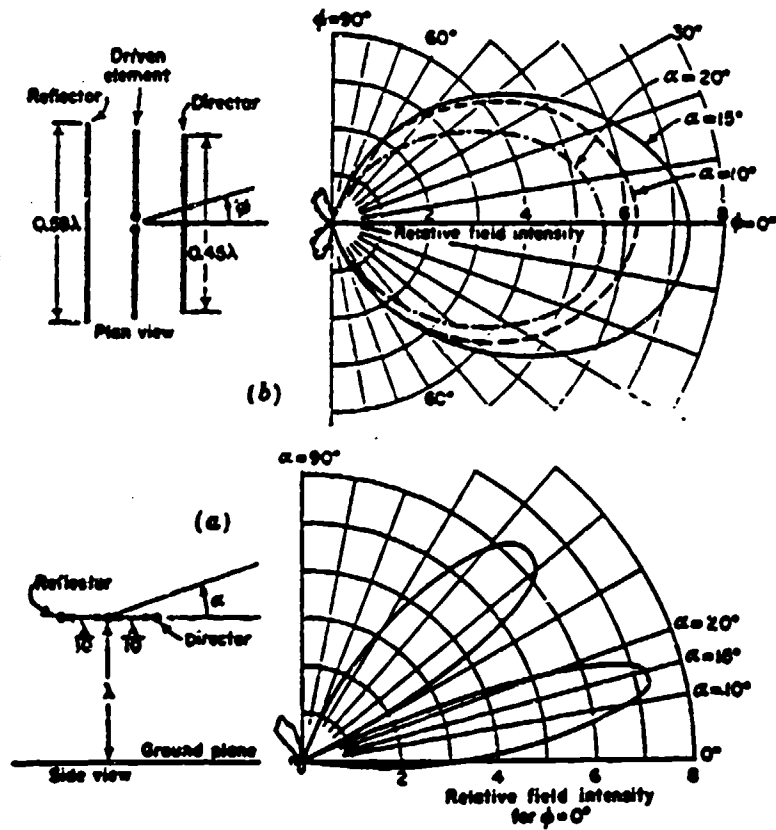
It has been found experimentally<sup>29</sup> that little is gained by adding more reflectors to the Yagi antenna, but that the gain may be increased by adding a large number of directors. An example, for an element spacing of  $0.34\lambda$ , is shown in Table 4.

---

<sup>27</sup> Actually the Yagi-Uda antenna, named after the two Japanese scientists who invented it.

<sup>28</sup> If the element spacing were increased from 0.12 to 0.452 (the optimum value for this antenna), the gain would be about 7 or 8dB.

<sup>29</sup> Jasik, Antenna Engineers' Handbook, pgs 5-24ff.



Source: Reference 7.

Figure 25. Three Element Yagi Antenna

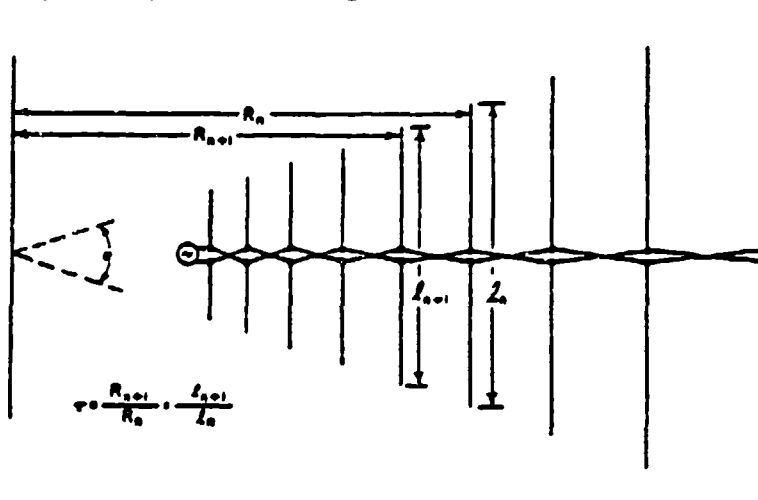
Table 4. Gains of Yagi Antennas

Number of Directors	Beamwidth (degrees)	Power gain over Dipole (dB)
30	22	14
20	26	13
13	31	12
9	37	11
4	46	9

The longer antennas are often used in the UHF and VHF bands, but are not suitable at HF because of their size. An array with a half-wavelength driven element and 20 directors at a frequency of 10 MHz, with an element spacing of  $0.34\lambda$ , would be 15 m wide and 215 m long. For this reason, HF Yagi antennas are generally limited to at most two directors.

Yagis are often used because they are simple and easy to install. They are usually mounted on a tower and can be rotated to point the main beam in the desired direction. The major limitation of the Yagi is its narrow bandwidth. Since the lengths of the reflector and director(s) (in wavelengths) is critical to their proper operation, changing frequency will detune them and spoil the main beam. Bandwidths of two percent are typical.

(9) **Log Periodic Antennas.** In this class of antenna, which is used when a broad tuning bandwidth is required,<sup>30</sup> the design of the structure is such that the electrical properties of the antenna repeat periodically with the logarithm of the frequency. A typical log periodic dipole array is shown in Fig. 26.



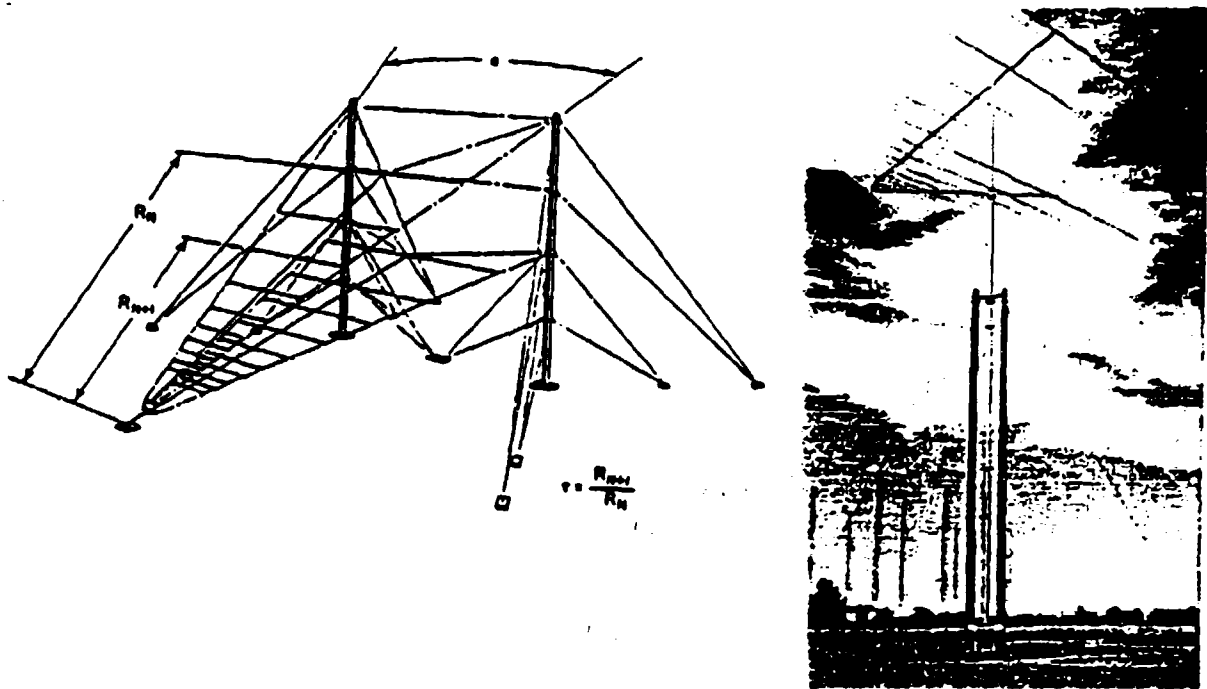
Source: Reference 5.

Figure 26. A Log-Periodic Dipole Array

The logarithmic requirement is satisfied by the fact that the ratios of element lengths and positions form a geometric sequence with the same ratio ( $\tau$  in Fig. 26).

Mast-mounted and fixed log periodic antennas for HF applications are shown in Fig. 27.

<sup>30</sup> For example, most home television antennas are of the log periodic type.



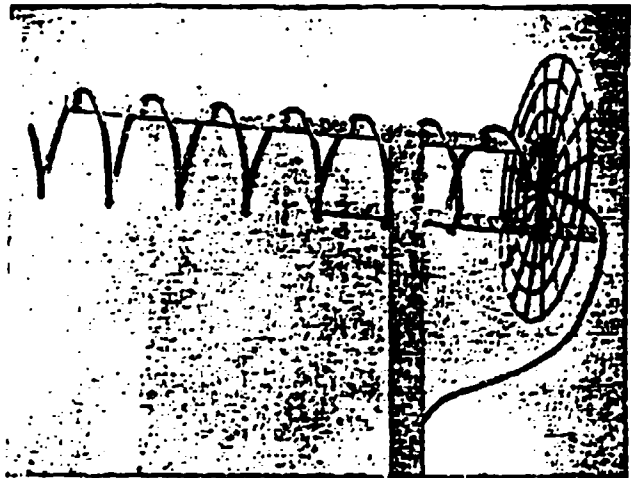
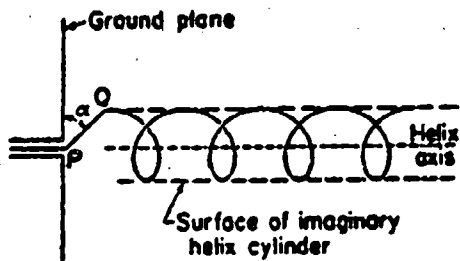
Source: Reference 5.

**Figure 27. HF Log Periodic Antennas**

The gain of a log periodic antenna like those in Fig. 27 will be about 10 dB over that of an isotropic antenna. The tuning range extends from the frequency at which the longest element is a half-wavelength long to the frequency at which the shortest element is a half-wavelength long, and can easily be made 10:1 or greater.

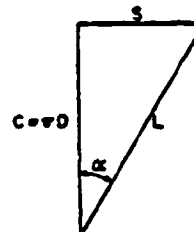
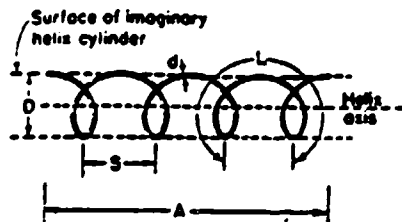
(10) **Helical Antennas.** This class of antenna has several unusual properties: It operates in an end-fire mode, produces waves that are circularly polarized, and maintains its performance characteristics over a frequency range of about 1.7 to 1.

A photograph and a schematic drawing of a typical helical antenna are shown in Fig. 28, and the pertinent parameters are shown in Fig. 29.



Source: Reference 7.

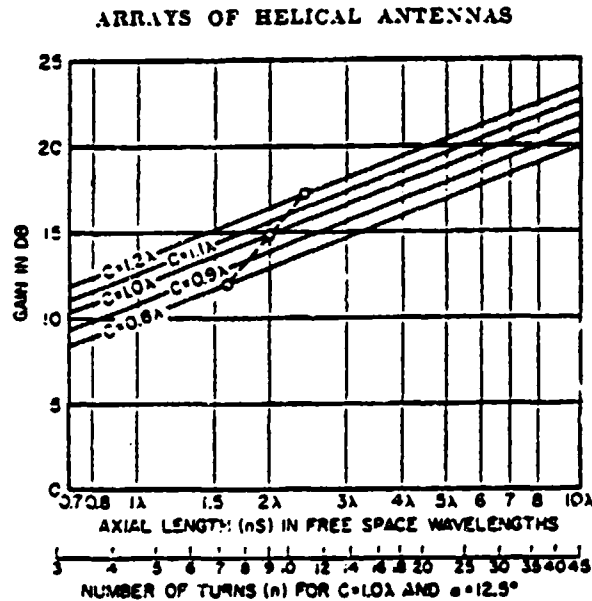
Figure 28. Helical Antenna



Source: Reference 7.

Figure 29. Parameters of Helical Antenna

Typically, the circumference of the helix is made equal to one wavelength ( $C=\lambda$ ), and the pitch angle  $\alpha$  is  $12.5^\circ$ . The gain of such an antenna as a function of its total length is shown in Fig. 30.



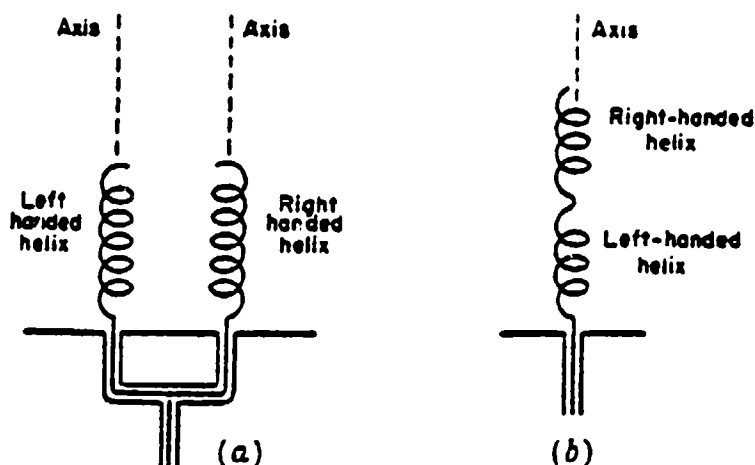
Source: Reference 5.

**Figure 30. Power Gain of Helical Beam Antenna with Respect to Isotropic Circularly Polarized Source**

An antenna with a circumference of one wavelength and a length of two wavelengths would have nine turns and a gain of 15 dB. At a frequency of 10 MHz ( $\lambda = 30\text{m}$ ), this antenna would have a diameter of 9.55m and a length of 60m.

The circularly-polarized wave from a helical antenna would be perfectly suitable for skywave path since, as noted earlier, polarization is unimportant in this case. Over a ground-wave path, half of the radiated power (the horizontal component of the circularly polarized wave) would be absorbed by the ground.<sup>31</sup> For this case, linear polarization can be achieved, at the expense of increased size and complexity, by using two helical antennas. They may be mounted either coaxially or side by side as shown in Fig. 31.

<sup>31</sup> This is a loss of 3 dB, which leaves a net antenna gain of 12 dB, which is still equal to or greater than the gains of other comparable HF antennas. See the earlier discussion of rhombic, yagi, and log periodic antennas.



Source: Reference 7.

**Figure 31. Helical Antenna Arrangements to Produce Linear Polarization**

## **I. TACTICAL HF ANTENNAS**

Some of the antennas described above are best suited for fixed or semi-permanent sites because they are large, awkward to erect, and do not disassemble into pieces small enough for easy transportation. This would include the helical, the log periodic and, probably, the Yagi.

A "Tactical HF Antenna" could be defined as one which can be easily assembled in the field by a few men, using simple components which can be carried in standard military vehicles: metal antenna masts, wire, insulators, etc. Antennas which satisfy this definition are:

- Single vertical quarter-wave monopoles
- Dipoles, both horizontal and vertical
- Arrays of vertical monopoles, phased for either broadside or end-fire operation. A screen of grounded vertical elements can be placed behind the broadside array to improve its performance.
- V antennas
- Rhombics



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The gains of these antennas (with respect to an isotropic antenna) are shown in Table 5. For consistency, the length of the array, and the leg lengths of the V and rhombic, were all chosen to be two wavelengths.

Table 5. Gains of Tactical HF Antennas

Type	Gain Over Isotropic
Vertical Quarter-Wave Monopole	5.2 dB
Half-Wave Dipole: Free Space	2.2 dB
Near Ground	5.2 dB
Array of Monopoles (2λ)	11 dB
- With Back Screen	14 dB
V Antenna (2λ)	11 dB
Rhombic (2λ)	13 dB

Using one of these antennas,<sup>32</sup> it should be possible to satisfy simultaneously the goals of placing a beam in the desired direction, placing a null in the direction of any source of interference, and establishing the proper vertical take-off angle for sky-wave circuits.

### J. SPREAD-SPECTRUM MODULATION

The last term to be discussed from Eq. 16 is B, the bandwidth filter factor. For signals with flat frequency spectra, such as the spread-spectrum signals to be discussed here, B was defined as:

$$B = B_I/B_R \quad (22)$$

where:  $B_I$  = Bandwidth of interfering signal

$B_R$  = Bandwidth of bandpass filter in receiver

The S/I ratio and, therefore, the performance of the radio link will be maximized by making B as large as possible. This can be accomplished in several ways:

---

<sup>32</sup> Some suggestions for simplified versions of these antennas may be found in FM 24-18 Tactical Single-Channel Radio Communications Techniques.

- The bandwidth of the receiver bandpass filter,  $B_R$ , should be made as narrow as possible while still passing all of the energy in the desired signal. This by itself will be effective against natural noise and certain sources of man-made noise (e.g. electrical machinery, corona discharges, etc.) which are inherently wide band sources.
- In the case of deliberate jamming, the goal is to force the jammer to increase the bandwidth of his signal while, at the same time, keeping the receiver bandwidth constant. This is the purpose of spread-spectrum modulation.

A typical HF signal using conventional modulation techniques (AM, SSB, FSK, or narrow-band FM) is limited to a channel bandwidth of 3 kHz, both by the propagation characteristics of the HF band and by the extreme crowding of the band. An enemy who wants to jam the signal can concentrate all of his jammer transmitter power into the narrow bandwidth of the desired signal.

Antijam modulation techniques operate by spreading the desired signal over a wider bandwidth, using some prearranged coding system so that it can be collapsed back to its original bandwidth at the receiver. The two most common techniques, frequency hopping and direct-sequence spread spectrum, will be discussed later.

When this is done, the jammer is forced to spread his power over the wider bandwidth. At the receiver, the desired signal will be collapsed back to its original bandwidth, but the jamming signal will remain spread because it is not modulated with the proper coded waveform. The antijam processing gain resulting from this process is given by the ratio of the spread bandwidth (RF bandwidth) to the signal bandwidth--that is, by the bandwidth filter factor. Some typical values are shown in Table 6.

**Table 6. AJ Processing Gain (dB)**

Signal Bandwidth	RF Bandwidth		
	500 kHz	5 MHz	15 MHz
3 kHz	22	32	37
6 kHz	19	29	34
12 kHz	16	26	31

The effectiveness of the jamming signal will be reduced by an amount equal to the Processing Gain (PG). The improvement in the signal-to-interference ratio caused by spread spectrum processing is given by:

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$$(S/I)_0 = S/I + PG \text{ (dB)} \quad (23)$$

where:  $S/I$  = on-the-air  $S/I$ , prior to processing  
 $(S/I)_0$  = signal-to-interference ratio after processing

It is clear from the preceding discussion that antijam effectiveness is maximized by spreading the signal over as wide a bandwidth as possible. In the HF band, however, this will be limited both by the availability of frequency assignments and by the propagation characteristics of the band itself.<sup>33</sup> Most current HF antijam systems spread over a bandwidth of only  $f_c \pm 2.5$  percent or less (a bandwidth of 500 kHz at a carrier frequency,  $f_c$ , of 10 MHz), and all HF systems which use sky wave propagation are constrained to operate between the MUF and the LUF. Fig. 8 shows that the difference between these two frequencies may be 2 MHz or less at night, and not more than 15 MHz during the day.

a. **Frequency Hopping.** Frequency hopping consists of a periodic changing of the transmitter frequency in a pattern determined by a pseudorandom sequence.<sup>34</sup>

It is classified as:

- Fast Frequency Hopping if there is at least one new frequency for each transmitted symbol (bit).
- Slow Frequency Hopping if two or more symbols are transmitted during each hop.

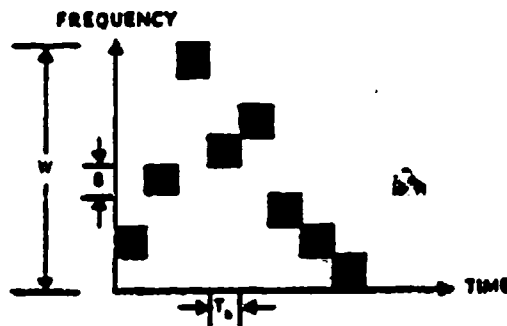
The simplest case, in which a single carrier frequency and data channel are used for each hop, is called single-channel modulation, and is shown schematically in Fig. 32,<sup>35</sup> where  $B$  is the channel bandwidth,  $W$  is the total RF bandwidth used by the frequency-hopping system, and  $T_h$  is the hop period.

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<sup>33</sup> For a discussion of HF propagation, see IDA Paper P-2170, *Radio Communication in the High Frequency Band*, UNCLASSIFIED.

<sup>34</sup> A pseudorandom sequence is a precisely defined, deterministic sequence of bits, usually generated by a digital shift register with a set of feedback loops which determine the sequence of bits. The sequence is easily matched and synchronized (decoded) by a receiver that knows the coding algorithm (the pattern of the feedback loops), but appears as random noise to a receiver that does not. The pseudorandom bit stream may also be encrypted if it is necessary to provide security as well as AJ and LPI.

<sup>35</sup> Since there is one hop for each symbol, this is fast frequency hopping.

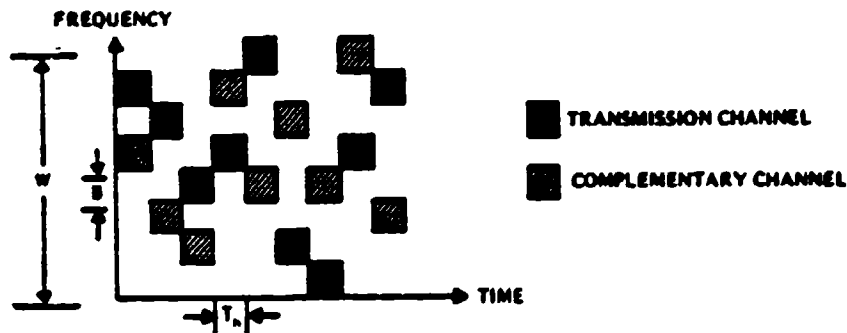


Source: Reference 13.

**Figure 32. Single Channel Fast Frequency Hopped Signal**

When binary FSK modulation is used, separate frequencies are assigned to represent the binary "1" (mark) and "0" (space). The pair of possible frequencies changes with each hop, and the one that is actually sent is called the transmission channel, while the one that would have been used if the other symbol were transmitted is called the complementary channel. This is easily extended to m-ary modulation, in which m frequencies are available on each hop to represent the m possible data symbols.

The hopping pattern for binary FSK is shown schematically in Fig. 33.



Source: Reference 13.

**Figure 33. FSK Fast Frequency Hopped Signal**

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At the receiver, a code generator identical to that at the transmitter<sup>36</sup> is used to drive a local oscillator (frequency synthesizer) in such a way that the frequency hopped signals are converted to a common and constant intermediate frequency. The process is called "dehopping" and is possible only for a receiver with the proper pseudonoise (or crypto) code.

At any given instant of time all of the power of a frequency-hopping transmitter will be concentrated on a single RF channel. A conventional communications or intercept receiver will detect the signal as an occasional burst of noise whenever the frequency-hopping pattern happens to land on the frequency to which it is tuned. However, a wideband (e.g., microscan) receiver scanning over the HF band will be able to detect the signal easily and unambiguously.

Frequency hopping reduces the effectiveness of a conventional wideband noise jammer by forcing it to spread its power over the whole hopping bandwidth ( $W$  in Figs. 32 and 33). If the actual channel bandwidth is  $B$ , the jammer power is thus diluted by a factor  $B/W$  with respect to the transmitted signal. This value, which is the AJ Processing gain, is shown in Table 6.

Another possible countermeasure against frequency hopping is to use a repeater jammer (or frequency-following jammer). A wideband receiver monitors the hopping spectrum and, each time a new hopping frequency is detected, a highly agile transmitter is tuned to that frequency and a jamming signal is transmitted. Although technically feasible and theoretically attractive, the technique has several practical limitations. If the RF band is crowded (as the HF band usually is), and especially if several frequency-hopping transmitters are operating simultaneously, the jammer will have difficulty deciding which signal belongs to the system it is trying to jam. It will have to do some rapid, and perhaps inaccurate signal sorting, and/or dilute its power in order to jam the several most likely candidates. In addition, there is a geometrical limitation that results from the fact that the repeater jammer must intercept the hopped signal, process it, tune its transmitter, and send the jamming signal, all while the victim receiver is still tuned to the same frequency.<sup>37</sup>

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<sup>36</sup> The process of obtaining and maintaining time synchronization between the two code generators is crucial, but will not be discussed in this paper. Suffice it to say that it can be done.

<sup>37</sup> See Torrieri [Ref. 13, pg. 159] where it is shown that the jammer must be located inside an ellipse that has the friendly transmitter and receiver as foci. For the low hop rates now used in the HF band, the ellipse is so large that this is not a serious limitation.

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b. **Direct-Sequence Spread Spectrum.** If an RF carrier with frequency  $f_c$  is phase-shift modulated by a binary random or pseudorandom sequence (i.e., if the phase of the carrier is shifted by  $180^\circ$  every time the sequence shifts from 1 to 0 or vice versa),<sup>38</sup> the resulting signal will be essentially identical to band-limited white Gaussian noise, and will have the frequency spectrum shown in Fig. 34.

The null-to-null bandwidth of the central peak of the spectrum will be twice the frequency at which the phase alterations are made, or  $2/T$  where  $T$  is the period of the modulating binary sequence. Thus, for example, modulation by a 1 Mbps pseudorandom sequence will spread the RF energy over a bandwidth of 2 MHz.

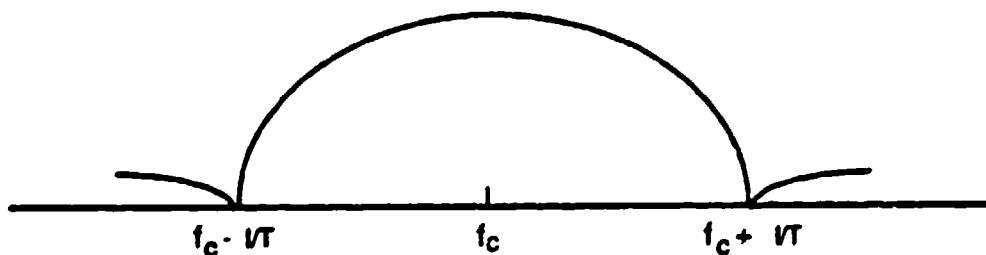
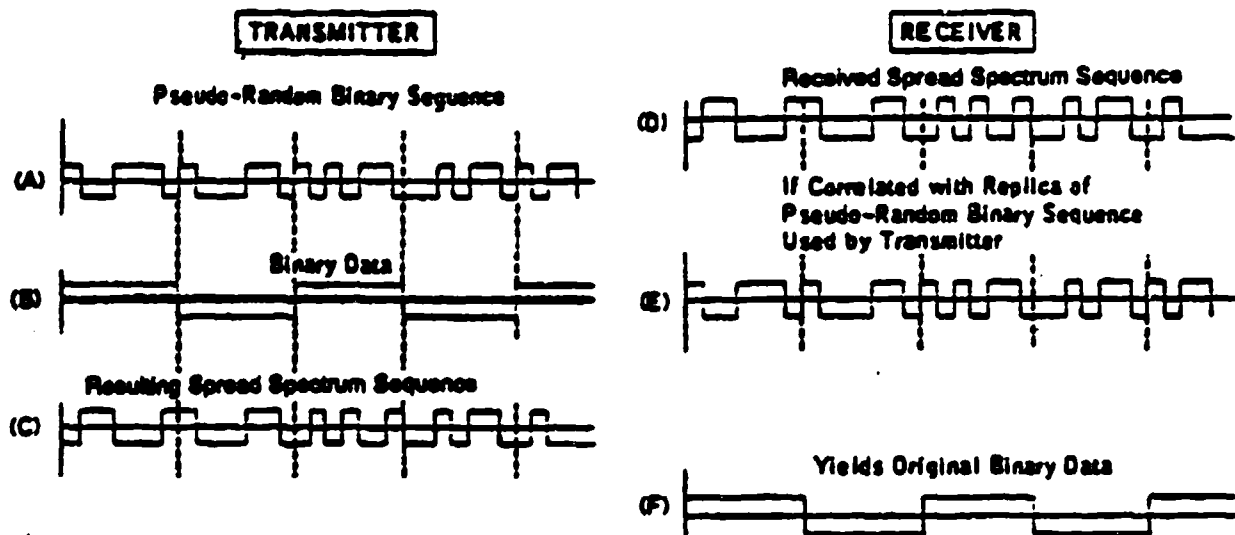


Figure 34. Spectrum of PSK Spread-Spectrum Signal

In direct-sequence spread spectrum, the binary data signal is added<sup>39</sup> to a pseudonoise binary sequence to form the spread-spectrum sequence which is used to PSK modulate the RF carrier. At the receiver, a replica of the pseudorandom binary sequence is subtracted from the received spread spectrum signal to yield the original binary data signal. This is shown in Fig. 35.

<sup>38</sup> Other modulation techniques, some of which are quite ingenious, have been used, but phase-shift modulation (or phase-shift keying) is the simplest and most common.

<sup>39</sup> This is Modulo-2 binary addition which has the following rules:  $0+0=0$ ,  $0+1=1$ ;  $1+1=0$ .



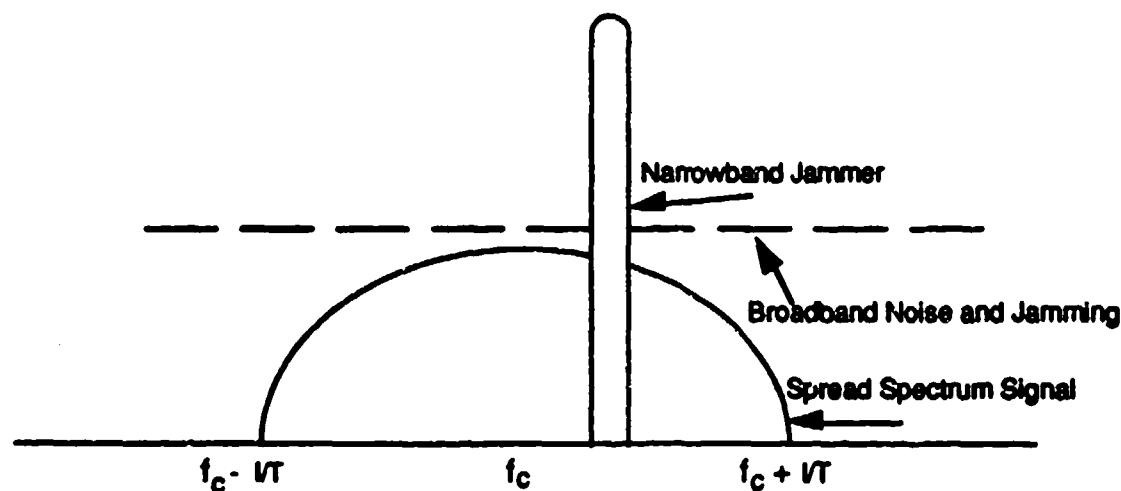
Source: Aviation Week & Space Technology, January 3, 1983.

**Figure 35. Direct-Sequence Spread Spectrum**

If the frequency of the pseudonoise sequence (called the chip rate) is considerably higher than the data rate (called the bit rate), as is invariably the case, the net result is to convert the data signal into a wideband noise-like signal. The processing gain (PG) or spreading ratio for this case is given by:

$$PG = \frac{\text{Spread Bandwidth}}{\text{Data Bandwidth}} = \frac{\text{Chip Rate}}{\text{Bit Rate}} \quad (24)$$

Since the transmitter power is spread over the entire RF bandwidth, the direct-sequence spread spectrum signal will, in many cases, have a lower amplitude than the background noise, interference, or broadband jamming. Fig. 34 shows such a signal in the presence of both broadband and narrowband jamming.



**Figure 36. Signal and Interference Before Despreading**

The despreading (decorrelation) process at the receiver will affect each of these signals differently:

- The original data signal will be despread and recovered. Its bandwidth will be reduced to its original value, and its amplitude will be increased from its spread value by an amount equal to the processing gain.
- The narrow band jamming signal will be spread, by exactly the same process that caused the spreading of the original data signal at the transmitter. Its amplitude will be reduced by an amount equal to the processing gain and its spectrum will become that of band-limited white noise.
- The broadband noise and jamming will remain essentially the same.<sup>40</sup> The actual signal waveform will, of course, change but the spectrum will remain white Gaussian noise.

The net result of the despreading process will be to recover the desired signal from the broadband noise and jamming, and to suppress the narrowband jamming, as shown in Fig. 37.

<sup>40</sup> This is because, mathematically, the correlation of a Gaussian signal (the noise and jamming) with another Gaussian signal (the pseudonoise sequence) yields yet another Gaussian signal.



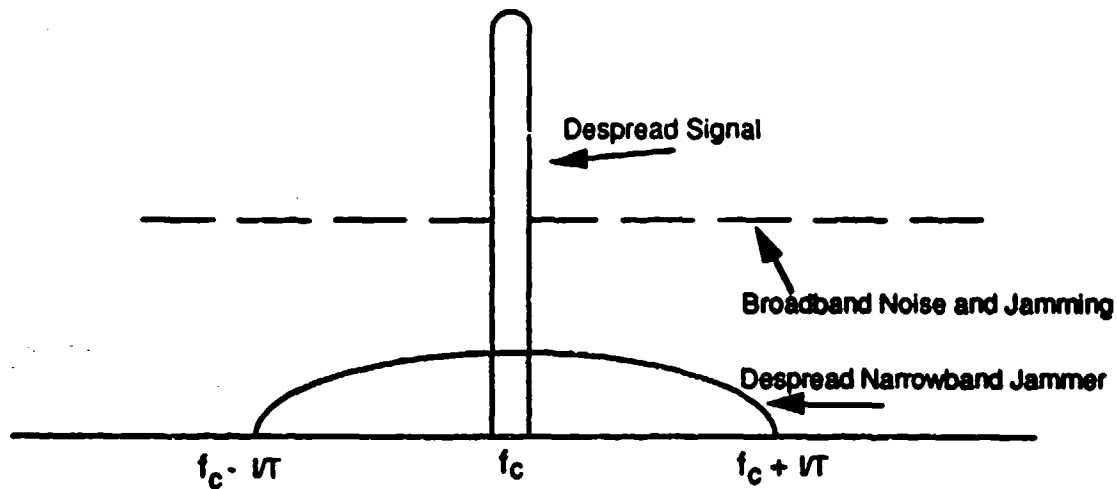


Figure 37. Signal and Interference After Despreading

Since the energy of a direct-sequence spread spectrum signal is distributed continuously across the entire spread bandwidth, it is very difficult to detect. It will appear only as a slight (usually insignificant) increase in the background noise level.

Narrowband and broadband jamming will be equally effective against direct-sequence spread spectrum since, as shown in the previous paragraph, both of these jamming signals will be converted to white noise by the despreading process at the receiver. For successful jamming, the jammer power must be sufficient to overcome the processing gain of the spread-spectrum system (see Eq. 24), and this usually requires an inconveniently large jammer transmitter.

c. **Comparison of Spread-Spectrum Systems.** The following points compare the features of frequency-hopping (FH) and direct-sequence (DS) spread spectrum. The list is not complete, and the technical arguments are not rigorous, but it should be clear that the present trend toward using FH should be carefully considered and possibly reversed.

- **Low Probability of Intercept/Low Probability of Detection (LPI/LPD).** It was noted earlier that, at any given instance of time, an FH transmitter has all of its power concentrated on a single narrow RF channel. A DS signal, on the other hand, spreads its power over a wide RF bandwidth. It will look like noise and may actually be buried in the ambient background noise. Thus, the FH signal can be detected by any HF receiver, and almost instantaneously by a rapid-scan (microscan) receiver, while the DS signal will be detected only if it is strong enough to raise the background noise level appreciably.

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- **Co-site Interference.** Mutual interference occurs when energy from one transmitter (including sidebands and harmonics) falls into the passband of a receiver operating in another net. The problem is particularly critical when the transmitter and receiver are at the same site (co-site interference), and is exacerbated when one or more of the nets at the site uses FH. Although it is possible to select and use non-interfering (orthogonal) hop sets, this places an additional burden on the frequency management system and, in any case, it is unlikely that enough channels will be available in the crowded HF band for such hop sets to be used. In the same situation, a DS signal will appear only as an increase in the background noise. In the worst case, a very strong DS transmitter may serve to reduce the sensitivity of receivers located close to it.
- **Frequency Assignment.** This may be a problem for either type of spread spectrum, but should be more manageable for DS. Frequency management agencies are reluctant to assign frequencies for FH systems, both because they are known to interfere with other radio equipment and because they require a large number of channels for effective ECCM operation. A few DS systems operating in the HF band should be acceptable, because they appear noise-like and will cause much less interference to other users. However, no precedent has been set for such frequency assignments, and the first few cases will probably be difficult.
- **Cost and Complexity.** A few years ago, when development of most of the current systems was started, FH was the preferred choice because the frequency synthesizers required for FH were readily available while the high-speed correlators required for DS were not. This is no longer the case and, for example, the Global Positioning System (GPS) will be using both 1 MHz and 10 MHz DS codes, and will be producing correlators inexpensive enough to be sold on the commercial market.
- **Acquisition and Synchronization.** Since the FH signal is strong and unique, while the DS signal is weak and noise-like, acquisition and synchronization are easier for FH systems.

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**Appendix A**  
**TASK BACKGROUND, OBJECTIVE, AND**  
**STATEMENT OF WORK**

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## APPENDIX A

### TASK ORDER BACKGROUND, OBJECTIVE, AND STATEMENT OF WORK

This IDA Paper was written in response to Task Order T-11-357 and Amendments 1 through 4. Those portions of the task order that pertain to the background, objectives, and statement of work are reprinted here.

#### 2. BACKGROUND:

The United States and the other NATO countries all have well-funded programs to develop anti-jam communications systems capable of overcoming the electronic countermeasures (ECM) threat posed by the Soviet Union and other potential enemies. In the United States, these programs include a new combat net radio (SINCGARS); data systems--the Joint Tactical Information Distribution System (JTIDS) and the Enhanced Position Locating and Reporting System (EPLRS); and High-Frequency Anti-Jam (HF/AJ) systems. Unfortunately, each NATO nation, and often each military Service of the nation, has its own program. These have been managed separately and apparently in isolation, with the result that the systems are not interoperable. For example, the U.S. SINCGARS is not interoperable with the German combat net radio (FUGER-A), and at one time, the U.S. alone had seven separate HF/AJ programs, no two of which could talk to each other.

Since interoperability is a prime requirement in combat operations, these programs must be studied to determine the means of achieving systems' interoperability within the same frequency band.

#### 3. OBJECTIVE:

The purpose of this task is to provide technical and analytical assistance and products to the Joint Staff in its task of ensuring that common interoperable electronic counter-countermeasures (ECCM) waveforms are developed for the communications systems of the U.S. military Services and our NATO allies.

#### 4. STATEMENT OF WORK:

**Specific Tasks and Deliverables.** This is a multi-year effort. While all open issues will be addressed, the effort will be focused on issues dealing with HF where the prospect for substantive results suitable as a basis for further Joint Staff action is greatest. The following tasks shall be addressed in FY 1991:

a. Continue preparation of a preliminary architecture study on HF to include identification of current and planned HF nets--the units/echelons/missions which participate in each net, identification of which nets need to interoperate, the type of traffic passed in each net, minimum essential nodes and rates required by C2 systems and supporting HF system nets, identification of the features needed for each network and the standards which apply (to include the need for ECM protection), the degree of Federal Agency, Joint Service, or Allied participation, and an analysis of how these features enhance/degrade the user's ability to circumvent frequency spectrum overcrowding, media, or threat conditions with which the user is confronted. Identify any radio nets, including

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strategic, tactical, and special operations, with special requirements which might mandate different ECCM standards. Special emphasis should be placed on HF ECCM requirements.

b. Analyze alternative HF ECCM techniques, compare capabilities, and recommend the minimum set of common AJ waveforms to satisfy the Requirement Submission for Advanced ECCM HF Radio Communications (RS JS 6-88).

c. Develop technical analyses of draft HF military and federal standards.

d. Provide technical analyses and descriptions of HF ECCM waveforms, and other inputs as required, to the developers of the TA/CE (Technical Analysis/Cost Estimate) in support of RS JS 6-88.

e. Analyze and compare the technical proposals and designs for ECCM systems provided by the U.S. military Services as well as those from domestic and foreign contractors, and make recommendations regarding feasibility, effectiveness, and interoperability. Competing factors such as available RF bandwidth, required AJ margin, data rate, and the operational environment will be considered in the analysis.

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**APPENDIX B**  
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